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AN EXPERIMENTAL RECEIVER FOR ULTRA-SHORT-WAVE RADIO-TELEPHONY WITH FREQUENCY MODULATION

by A. van WEEL.

621.396.5: 621.396.621

A description is given of a superheterodyne receiver for frequency-modulated signals, which forms part of an experimental ultra-short-wave radio telephony link on wave lengths in the neighbourhood of 1 m. This link was designed for the simultaneous transmission of 48 calls ("channels") and is therefore able to handle modulation frequencies up to about 200 kc/sec. The most interesting detail of the receiver is the push-pull mixing stage, equipped with triodes, which is made self-oscillating by introducing, in addition to the normal symmetric push-pull input circuit tuned to the signal frequency, an "asymmetric" input circuit tuned to the local oscillator frequency, which is generated in the same circuit as a consequence of the coupling between anode and grid circuits. As no separate oscillator valve is necessary, an appreciable part of the fluctuation noise is eliminated. As a consequence of the favourable properties of the mixing stage, as far as the noise is concerned, a high-frequency amplifier stage would give only little improvement in the ratio between intensity of signal and noise. The receiver, therefore, does not possess a high-frequency amplifier stage. The provisions for automatic volume control, usually found in receivers for amplitude-modulated signals, could also be omitted in this frequency-modulated receiver, thanks to the large suppression factor of the limiter.

A new transmitting-receiving apparatus has been developed for the experimental ultra-short-wave radio-telephonic link between the Philips factories in Eindhoven and those in Tilburg, which link has been in existence for a number of years already. After having described the transmitter in a previous article [1]), we shall now discuss the receiver. For the convenience of the reader we shall repeat briefly the most important facts about the installation. For the connection in one direction a wave length of 90.5 cm is used, in the other direction 99 cm. The transmitter and receiver function as link in a carrier telephony system with which 48 channels can be transmitted at the same time on one pair of conductors (in this case, on one radio wave). For this purpose both the transmitter and the receiver must be able to handle modulation frequencies up to about 200 kc/sec. For this link frequency modulation is employed. The maximum frequency swing, *i.e.* the largest deviation of the frequency emitted, compared with the average transmitter frequency (332.1 Mc/sec for one direction, 303.0 Mc/sec for the other), amounts to 0.6 Mc/sec.

General construction of the receiver

The receiver works on the superheterodyne principle. A block diagram of the most important parts is given in *fig. 1.* The signal received by the

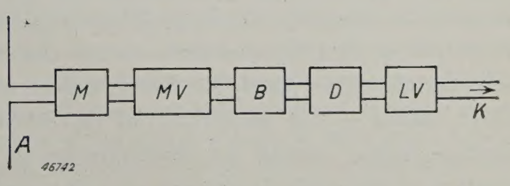

Fig. 1. Block diagram of the receiver; *A* aerial, *M* mixing stage, *MV* intermediate-frequency amplifier, *B* limiter, *D* frequency detector, *LV* low-frequency amplifier, *K* cable to the carrier-telephone apparatus.

aerial is applied to a mixing stage where the signal frequency is converted into an intermediate frequency of 18 Mc/sec. The converted signal is amplified and then passed through a limiter, the function of which is to suppress any amplitude modulation which may be present, especially the noise which

1) A. van Weel, Philips Techn. Rev. 8, 121, 1946.

occurs in that form [2]). The output of the limiter is detected and the low-frequency signal obtained is amplified and then passed through a cable to the apparatus for carrier telephony in the telephone exchange, where the 48 channels are split up and can be connected with the corresponding subscribers.

In principle this construction of the receiver does not differ from the normal receiver for frequency-modulated signals. However, it will be noted that there is no high-frequency amplifier stage preceding the mixing valve. Such a stage is in general desirable to improve the ratio between the intensity of the signal and that of the noise. In our case, however, the mixing stage has such good properties, as far as noise is concerned, that the addition of a high-frequency amplifier stage could have produced only a very small improvement. That stage could therefore be omitted, which meant a considerable simplification of the receiver.

In the following we shall discuss mainly the design of the mixing stage, while the other parts of the receiver will only be considered briefly.

Principle of the mixing stage

Mixing in triodes

The familiar multigrid mixing valves, generally employed for the conversion of signals on wavelengths longer than about 10 m, are no longer suitable for wave lengths of about 1 m, their special advantages being lost in this case because: 1) the shielding effect of the various screen grids has practically no result, since, due to the self-inductions and mutual inductions of the internal leads to these screen grids at these high frequencies, it is practically impossible to keep the grids free of high-frequency voltage variations; 2) the high internal resistance of the multigrid mixing valves, seen from the intermediate-frequency circuit, means no advantage because, to amplify the large frequency bandwidths required, heavily damped circuits have to be used. On the other hand the disadvantage of all multigrid valves — the occurrence of the so-called distribution noise caused by fluctuations in the distribution of the cathode current over the various current carrying grids — is the same with short waves as with longer waves. In the ultra short wave region, therefore, diodes or triodes are used as mixing valves.

The advantage of a diode mixing stage lies in the high input impedance, which makes a high step-up ratio of the signal delivered by the previous stage. However, the conversion amplification, *i.e.* the ratio between the intermediate output voltage and the high-frequency input voltage, cannot exceed unity.

When a triode is used as a mixer the input impedance is smaller than with a diode. On the other hand the conversion amplification is in general considerably larger than unity.

Further, it has been found from theoretical investigations [3]) that here a good ratio can be obtained between signal intensity and noise. These facts, together with another important advantage which will be discussed farther on, led us to employ triode mixing in the receiver in question.

Mixing in a triode is in principle accomplished by applying to the grid of the triode, together with the high-frequency signal voltage, a second voltage from a local oscillator (usually indicated briefly as the "local oscillator voltage").

The anode current then contains an intermediate-frequency component (difference of the frequencies of the two voltages mentioned), and by introducing in the anode circuit a circuit tuned to the intermediate frequency the desired intermediate-frequency output voltage is obtained.

At the high frequencies with which we are dealing here it is in general an advantage to construct amplifier and mixing stages on the push-pull principle [4]). The following general rule is then valid: of the three voltages occurring in a mixing stage, namely high-frequency signal voltage, local oscillator voltage and intermediate output voltage, two must always be in push-pull, while the third has to be in the same phase for the two mixing systems. Because of the symmetrical dipole aerial it is reasonable to apply the high-frequency signal in balance to the convertor valves. We now apply the equal-phased local oscillator voltage in the same sense to both valves and then obtain the intermediate frequency voltage in balance again. In that way we arrive at the diagram sketched in *fig. 2*, showing the principle of a push-pull mixing stage with two triodes. The connection of the anodes with the intermediate-frequency circuit and the source of anode voltage are omitted temporarily for the sake of simplicity. Between the control grids there is a circuit L_1-C_1 tuned to the signal frequency, which is inductively coupled with the aerial. In this way an e.m.f. in the aerial gives rise to grid a.c. voltages which are equal in magnitude

[2]) See for example Th. J. Weyers, Philips Techn. Rev. **8**, 42 and 89 1946.

[3]) See A. van Weel, thesis, Delft 1943.

[4]) See in connection with this and the following: M. J. O. Strutt and A. van der Ziel, The diode as a frequency-changing valve, especially with decimeter waves, Philips Techn. Rev. **6**, 285, 1941. In that article will be found explanations and the literature about the properties of valves on decimeter and meter waves, which have merely been mentioned in the above.

but opposite in phase for the two triodes (thus balanced). Between the middle of the circuit self-induction L_1 and the earth point (*i.e.* the chassis) there is an oscillator for the auxiliary voltage, so

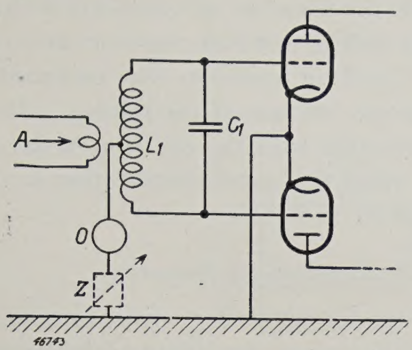

Fig. 2. The signal from the aerial is applied to the grids of the two triodes of the push-pull mixing stage by inductive coupling with a tuned push-pull input circuit (L_1-C_1). The auxiliary voltage (oscillator voltage) delivered by the local oscillator Q is applied to the grids in equal phase. The "asymmetric" circuit can be tuned to the oscillator frequency by an impedance L. The intermediate-frequency circuit to be connected between the anodes is here omitted for the sake of simplicity.

that the latter comes in equal phase between the grids of the two triodes and the cathode, which are connected with the earth point.

Separate tuning of the oscillator circuit

With such connections it is often found to be difficult to obtain a sufficiently high auxiliary voltage on the mixing valves, due to the fact that the asymmetrical circuit (earth point — oscillator — circuit self-induction — grid-cathode capacities of the valves — earth point) is not in general tuned to the oscillator frequency. This tuning can, however, easily be realized by including a tuning element somewhere in the circuit. Since the push-pull circuit, tuned to the signal frequency, may not be affected by this second tuning, the tuning element should be included in a part of the circuit through which no balanced currents flow. In fig. 2 the element in question is shown with a dotted line.

It is perhaps advisable to place some emphasis on the principle applied here: *to the input electrodes of two valves in push-pull connection two independent circuits tuned to different frequencies can be connected, viz. as push-pull circuit and an asymmetric circuit.* The separate tuning of each circuit can be varied with the help of circuit elements through which flow, respectively, only balanced currents or only equal-phased currents.

Excitation of the auxiliary voltage by the mixing connections themselves

In order to eliminate the objection of the above-mentioned low input impedance of the triodes at high frequencies, back-coupling is employed in the asymmetric circuit. This causes a reversal of damping in that circuit or, in other words, an increase in the input impedance of the valves and with it an increase of the auxiliary voltage obtained on the input electrodes.

If the reversal of the damping is made large enough (the back-coupling strong enough) the circuit itself begins to oscillate, and it oscillates at the frequency to which the asymmetric input circuit is tuned, *i.e.* the frequency of the auxiliary voltage.

A separate oscillator is then not needed for the excitation of this auxiliary voltage. As far as the conversion effect of the triodes is concerned it makes no difference whether the auxiliary voltage between grid and cathode comes from a separate oscillator or whether it is excited by the triodes themselves. The mixing action is determined exclusively by the magnitude of the auxiliary voltage and the non-linearity of the i_a-v_g characteristic.

The back-coupling mentioned in the case of a triode can be obtained in the familiar way by conecting a self-induction between the anode and the cathode.

This can be understood most easily by writing the equations for the scheme shown in *fig. 3*. If for the sake of simplicity we disregard the internal resistance of the valve these equations are as follows:

$$v_g = \frac{i_{g1}}{j\omega C_{ag}} - (i_a - i_{g1})\, j\omega L = \frac{i_{g2}}{j\omega C_{gk}},$$

$$i_a = S \cdot v_g.$$

From this it follows that:

$$\frac{i_g}{v_g} = \frac{i_{g1}}{v_g} + \frac{i_{g2}}{v_g} = j\omega C_{gk} + \frac{1}{j\omega L + \frac{1}{j\omega C_{ag}}} - \frac{\omega^2 SLC_{ag}}{1 - \omega^2 LC_{ag}}.$$

While with short-circuited self-induction L the triode, as an element of the input circuit, can be replaced by the capacities C_{gk} and C_{ag} connected in parallel, it may be seen from the last equation that by the intermediate connection of the self-induction the equivalent connections become a connection in parallel of C_{gk} with L and C_{ag} in series and a negative resistance. The latter causes the desired increase in the input impedance of the valve, since the admittance i_g/v_g is lowered by that term.

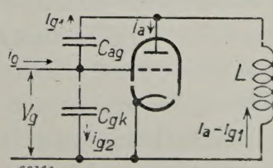

Fig. 3. Back-coupling of a triode *via* the grid-anode capacity C_{ag} and a self-induction L in the anode circuit of the triode. The total grid current i_g is composed of the two parts i_{g1} and i_{g2}, of which only i_{g1} is important for the back-coupling.

In the push-pull mixing stage the self-induction is introduced in the manner shown in principle in *fig. 4*. The equal-phased circuit, which is tuned to the desired oscillator frequency by means of the variable impedance Z, is now so strongly regenerated by the self-induction L_2 as to cause oscillation.

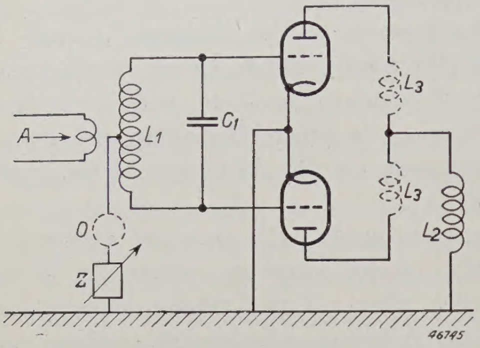

Fig. 4. The "asymmetric" input circuit of the push-pull mixing stage is regenerated by the self-induction L_2. The push-pull input circuit can be regenerated quite independently by the self-induction L_3.

It must be pointed out that the push-pull circuit can also be regenerated in a similar manner. For that purpose the two self-inductions L_3 shown with dotted lines in fig. 4 have to be placed in the anode circuit. In this way it is possible to obtain a stronger excitation of the signal delivered by the aerial, and thus a better relation between signal and noise [5]). The degree of back-coupling can be chosen different for the push-pull circuit and for the asymmetric circuit, but L_2 has no effect on the push-pull circuit. This is important because the back-coupling for the push-pull circuit may not, of course, be made so great that oscillation (with signal frequency) occurs in that circuit.

The ratio of signal intensity to noise

The fact that a separate oscillator valve can be omitted constitutes of itself a welcome simplification of the connections. Still more important is the fact that an improvement in the relation between signal intensity and noise is thereby obtained. This can easily be explained. In the anode current of an oscillator valve, as is the case with every valve, there is a noise with a continuous frequency spectrum. The circuit connected with the anode circuit of the oscillator valve, which is tuned to the oscillator frequency, acts as a kind of filter which passes chiefly the oscillator frequency but to a certain extent also the neighbouring signal frequency and, to a smaller extent, the much lower intermediate frequency. The components of the noise with those frequencies are thus also applied more or less intensely to the mixing valve as input voltage together with the oscillator voltage proper. The noise at signal frequency is then also converted with the signal, the noise at intermediate frequency is directly amplified, and both therefore give rise to an increased level of noise in the intermediate-frequency output voltage of the mixing valve.

It is clear that with the omission of the separate oscillator valve this extra contribution to the noise is eliminated.

Choice of the intermediate frequency

The above-described effect of the noise of an oscillator valve is stronger according as the intermediate frequency chosen is lower. This means that the signal frequency lies very close to the oscillator frequency, so that the component of the noise at the signal frequency is passed by the tuned circuit of the oscillator almost in full strength. From this point of view, therefore, the intermediate frequency should be as high as possible.

Although in our case, due to the omission of the oscillator valve and the noises inherent therein, this consideration is no longer applicable, as far as the choice of the intermediate frequency is concerned one reaches the same conclusion. In the above considerations it was tacitly assumed that the symmetry of the push-pull stage is complete, so that the balanced and asymmetric circuits do not affect each other at all. In practice there will always be a certain asymmetry and consequently a coupling between the two circuits. If, now, the characteristic frequencies of the two circuits lie close together (*i.e.* if the intermediate frequency is low), the oscillation of one circuit causes resonances in the other, with the further consequence of unequal voltages on the two valves, etc. The intermediate frequency may not, therefore, be chosen too low. Upon closer consideration it is found that it should amount at least to 1/20 of the signal frequency in order to render the coupling in question harmless. The receiver described here works with an intermediate frequency of 18 Mc/sec; the condition just given is therefore satisfied up to a signal frequency of 360 Mc/sec, *i.e.* a wave length of 83 cm.

Complete connections of the mixing stage

In *fig. 5* a diagram is given of the complete connections of the mixing stage, while *fig. 6* shows a photograph of the stage when constructed.

To the grids of the two button triodes used as mixing valves a Lecher system is connected, which together with a condenser (C_1) forms the

[5]) A. van Weel, loc. cit.

push-pull input circuit of the mixing stage. The capacities C_2 serve only for separating the D.C. voltage situation of the two grids. The desired regeneration of the push-pull input circuit is obtained by the self-induction L_3 introduced between the anodes.

self-induction, however, not only determines the degree of back-coupling but also has an influence on the frequency generated. This may be seen from the equation derived in connection with fig. 3, which indicates that L_2 introduces not only a negative resistance but also a reactive component in the

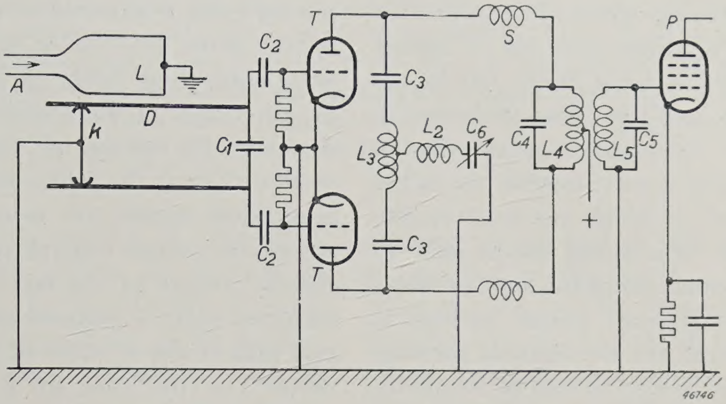

Fig. 5. Diagram of complete connections of the mixing stage. T triodes, D Lecher system with sliding short-circuiting bridge K and tuning condenser C_1, A aerial connection with coupling loop L; L_2 and C_6 back-coupling elements of the "asymmetric" input circuit, in which C_6 also serves for the fine tuning to the local oscillator frequency, L_3 back-coupling element of the push-pull input circuit, C_4-L_4-L_5-C_5 band filter for the intermediate frequency amplifier. S high-frequency choke coils, C_2, C_3 coupling condensers.

The asymmetric input circuit of the mixing stage is formed by the two conductors of the Lecher system, which for this purpose are to be considered simply as two separate impedances through which rectified currents pass. This input circuit is closed by the chassis, since the sliding short-circuiting bridge of the Lecher system is connected directly with the chassis. The back-coupling for the equiphased oscillation is accomplished by the self-induction L_2 connected to the middle of L_3 (to which has to be added the two halves of L_3 to be considered as connected in parallel). The magnitude of this

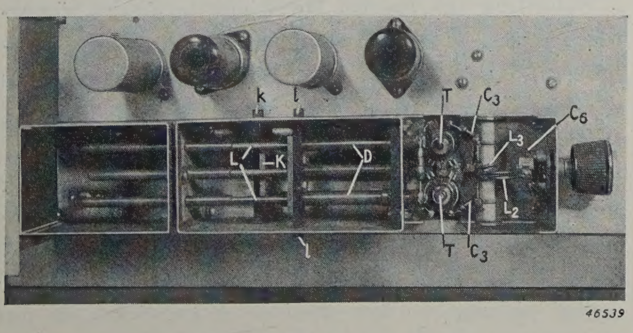

Fig. 6. The push-pull mixing stage with shielding cap removed, showing the two button triodes T and the input Lecher system D, the short-circuiting bridge K of which can be moved by means of the projections k. In front of that system may be seen the "loop" L, which can be varied with the projections l and which makes the coupling with the aerial cable and whose middle point remains in connection with the earth point (the chassis) by way of the middle conductor. To the right of the button triodes may be seen the capacities and self-inductions indicated in fig. 5, C_3, C_6, L_3, L_2.

equivalent connections for the valve. Since the adjustment of the back-coupling is not very critical, advantage can be taken of the influence mentioned to regulate the oscillator frequency more exactly by varying L_2. Since, however, a continuously variable self-induction is less easy to realize, for this purpose a rotating condenser (C_6) is introduced in series with L_2.

The tuning of the push-pull and asymmetric circuits is now accomplished as follows. After L_2 and C_6 have been so chosen that sufficient regeneration has been obtained for the occurrence of oscillation in the asymmetric circuit, that circuit is tuned approximately to the oscillator frequency, by moving the short-circuiting bridge of the Lecher system. By a correct choice of C_1 the push-pull circuit is then tuned approximately to the signal frequency, while the fine tuning is obtained again by a slight movement of the short-circuiting bridge, and the fine tuning to the oscillator frequency is finally obtained with C_6. This rather laborious manipulation constitutes no objection in our case, since the receiver always remains set on the same wave length. Finally, there is the intermediate-frequency circuit. This consists of the self-induction L_4 and the capacity C_4. Between the triodes and this circuit the high-frequency choke coils S are connected in order to prevent the high-frequency circuits from being affected by L_4-C_4. Conversely, the capacities C_3,

which have a high impedance for the intermediate frequency and a low impedance for the signal and oscillator frequency, provide that the circuit L_4-C_4 is not practically short-circuited by the small self-induction L_S.

Remaining parts of the receiver

The intermediate-frequency amplifier consists of a number of stages with EF 51 valves. The valves are coupled by band filters, but in one stage connections are used like those described in the article about the transmitter of this installation [1]). By this means it was possible (as in the case of the transmitter) to divide the receiver into two parts in this stage, thus at the middle point of the intermediate-frequency amplifier, and to house the two parts in two separate panels without it being necessary to shield the connections between the panels. The two panels may be seen in the photograph of *fig.* 7.

After the intermediate-frequency amplification a limiter is applied to the signal, as is customary in receivers for frequency-modulated signals. This consists of two valves provided with grid condenser and leak resistance. When the amplitude of the A.C. voltage on the grid of a valve in these connections increases, the grid becomes more negative due to the occurrence of grid detection. This reduces the anode current and, in spite of the increased input voltage, the output voltage of the valve shows little or no increase. In this way any amplitude modulation which may still be present in the frequency-modulated signals, and especially the noise present in that form, is suppressed to a large extent. The output voltage of the first limiter valve, which is already quite constant, is chosen so high that the second valve is adjusted at the point of the characteristic most favourable for the limiting effect. A total suppression factor of 100 is thereby attained, *i.e.* the depth of the amplitude modulation of the signal at the output of the limiter is 100 times smaller than at the input. Receivers for amplitude-modulated signals are usually equipped with an automatic volume control, with which a part of the output voltage of the last intermediate-frequency amplifier valve is rectified and used to regulate the grid bias of one or more of the preceding amplifier valves. In the case of a frequency-modulation receiver, when the limiter has a sufficiently large suppression factor such a volume control may be omitted, since the limiter already provides for a practically constant output voltage. In our case, therefore, all the valves of the intermediate-frequency amplifier work with constant negative grid bias. Account must then be taken, however, of the possibility that the last valve of the amplifier may be overloaded and damaged by a strong signal. In order to prevent this, that valve, as well as the valves of the limiter, is provided with a grid condenser and leak resistance in such a way that the valve simply begins of itself to act as a limiter as soon as the signal exceeds the permissible strength. The only result of an increase in the signal then is that there is a stronger suppression of any amplitude modulation which may be present, which at the most can but improve the quality of reception.

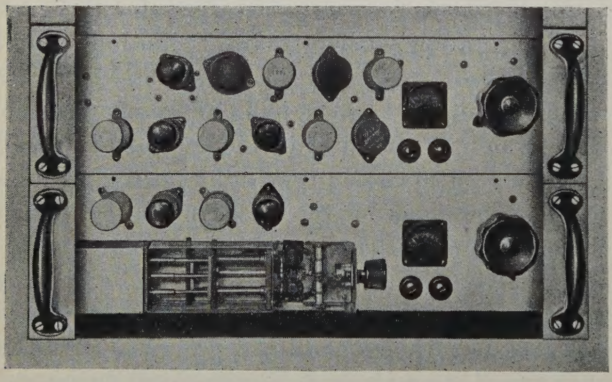

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Fig. 7. The receiver is assembled in two panels, which can be slid in and out of the bay containing the whole transmitter-receiver installation by means of handles. Upon sliding them in, all the necessary connections are made automatically by means of plugs on the panels and sockets on the bay. The meters with the adjacent switching knobs on each panel serve for checking the cathode current of all the valves.

Following the limiter there comes finally a normal frequency detector and a low-frequency amplifier valve. The low-frequency signal obtained is conducted *via* a transformer to the cable leading to the carrier-telephony apparatus.

After the previous description of the construction of the transmitter it is unnecessary to go further into that of the receiver. For several particulars reference is made to figures 6 and 7 and the annotation thereby.

HIGH-VOLTAGE RECTIFIER VALVES FOR X-RAY DIAGNOSTICS

by J. H. van der TUUK. 621.314.671:621.386.1:616.073

The high-voltage rectifier valves used in X-ray installations, especially in diagnostic apparatus, have to satisfy in the first place the following requirements: they should be resistant to high voltages, furnish the necessary large peak currents with a reproducible and not too large voltage loss, and should not develop too much heat. In this article it is explained in what way two fundamental types of valves, namely vacuum and gas-filled valves, can satisfy the requirements mentioned. Gas-filled valves are found to possess several properties which are very favourable for the object in view. However, also in the field of vacuum valves constructions have recently been developed which give interesting results, thanks to the use of thoriated tungsten cathodes, and which may supplant entirely or for a large part gas-filled valves because of their universal applicability.

In principle, after being stepped up, the alternating voltage from the mains can be applied directly to an X-ray tube. In the half cycle in which the hot cathode is negative and the anode positive X-rays are excited, while in the other half nothing happens. The fact that in many cases it is, nevertheless, found preferable to apply direct voltage or at least pulsating direct voltage to the X-ray tube has two reasons. In the first place when X-rays are excited in both halves of the cycle, with a given peak value of the tube current, double the X-ray intensity is obtained and, moreover, upon smoothing the rectified voltage X-ray output becomes larger, compared to the heat developed. In the second place, when an X-ray tube works on A.C. sometimes something does happen in the "other" half of the cycle, and what happens is extremely undesirable, *viz.* the so-called back-lash. This may occur in the half cycle in which the anode is negative and starts to emit electrons, either by becoming too hot (thermal emission) or because the electric field at its surface exceeds a certain value (cold emission) or it is struck by positive ions of gas. Since such a back-lash often leads to the destruction of the tube it must be prevented under all circumstances. When rectified voltage is applied the problem of back-lash, which presents special difficulties in the case of the X-ray tube because of the great heat development on the relatively small focus [1]), is transferred to the high-voltage valves, which serve for the rectification. These valves are very similar to the X-ray tube itself: they, too, possess a cathode and an anode in a vacuum-tight envelope. Since, however, in the case of valves the electrons carrying the current in the direction of transmission need not be strongly accelerated and focussed, local intense heating of the anode can be avoided by suitable construction, so that this important source of back-lash is eliminated.

Let us now consider somewhat more closely the requirements demanded of high-voltage rectifier valves. From the above follows the general requirement of voltage reliability, *i.e.* resistance to the high-voltage in the negative phase without danger of back-lash. In the positive phase, of course, the rectifier valve must be able to pass the current necessary for the X-ray tube. In the case of X-ray tubes for diagnostics it is, at present, a question of short-lived peak currents of 1 to 1.5 A, while maximum voltages up to 125 kV occur. In the case of therapy tubes much smaller currents are used, for instance maximum 40 mA continuously, with, however, considerably higher voltages, *viz.* 200 to 400 kV.

We shall devote ourselves here especially to the high-voltage rectifier valves used for diagnostics. In this case there are several additional requirements. For easy operation of an apparatus for diagnostics it is desirable that a certain tube voltage always corresponds to a definite position of the voltage regulator; for that purpose the voltage loss depending on the current should be small in all the parts of the high-voltage generator (transformer, valves, connections) and it should be as independent as possible of the circumstances of operation. The valves may, therefore, even at the highest currents, only require a low and reproducible voltage in the direction of transmission. This is especially true when the apparatus is provided with an automatic adjustment for the tube current [2]).

In the case of modern apparatus for diagnostics the valves are often housed in the same oil-filled

[1]) See the discussion of the problem of back-lash in the article J. H. van der Tuuk, Hard-glass X-ray tubes in oil, Philips Techn. Rev. 6, 309, 1941.

[2]) Such an arrangement, which serves to load up the focus of the X-ray tube automatically to the permissible temperature at every adjustment of tube voltage and loading time, is discussed in detail in: H. A. G. Hazeu and J. M. Ledeboer, A universal apparatus for X-ray diagnosis, Philips Techn. Rev. 6, 12, 1941.

container with oil as the high-voltage transformer, etc. Since, in order to keep the voltage loss (*i.e.* the copper and iron losses) constant, the whole generator may not become too warm, the additional requirement should be made of the valves that they develop little heat. On the one hand this is already ensured if the valve only takes up a low voltage in the direction of transmission, since the anode dissipation is proportional to it; on the other hand it also makes desirable a high specific emissivity of the cathode, since then only a small filament power is used for the emission of the peak currents.

We shall now examine several old and new valve constructions and ascertain how, in each case, the requirements mentioned can be satisfied.

Vacuum rectifier valves

Fig. 1 shows diagrammatically the construction of the old vacuum rectifiers as they were used for many years in all kinds of high-voltage installations. The envelope of the valve is so far evacuated

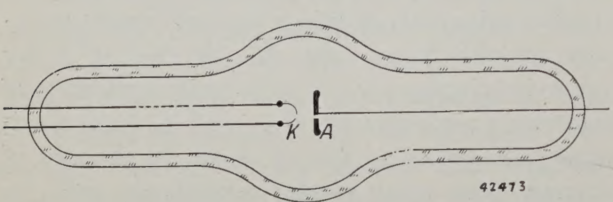


Fig. 1. Diagram of the cross-section of a vacuum rectifier valve as used formerly in high-voltage installations. An anode plate A is situated at a short distance opposite a tungsten cathode K in an evacuated glass tube.

that no breakdown can be caused by gas ions. The most important factor which should be taken into account to prevent back-lash is therefore the cold emission of the anode. In order to remove any such danger the field strength at all points on the surface of the anode should remain below a certain value, for instance 10^6 V/cm. For this purpose sufficient distance between the electrodes is necessary. Furthermore, the anode should be properly smooth-finished, in order to avoid local increases in the field strength, which are known to occur wherever there are slight irregularities in the surface. (In this respect also the problem of back-lash is easier to solve in the case of a rectifier valve than in that of an X-ray tube: in the latter case the anode surface at the position of the focus is always rendered slightly rough in the course of time by the intense heating (evaporation).)

As far as the distance between the electrodes is concerned, the larger it is made the greater the influence of the space charge on the variation of the field and the higher the voltages necessary in the direction of transmission to draw the desired current from the hot cathode. Therefore it is clear that the distance between the electrodes will not be made greater than absolutely necessary. For 100—125 kV, for example, a distance of 8—9 mm is sufficient. With a given valve according to fig. 1 the voltage loss amounted to about 2500 V for a current of 1A.

The small distance between electrodes, arrived at in this way, makes it impossible to use an oxide cathode, for there would be too great a danger of traces of barium from such a cathode striking the anode. Due to the increase in field strength at the unevenness thus formed (and due to the low work function of the barium) such a spot might act as a centre of cold emission in the counter-phase. Therefore in this type of valve tungsten cathodes are always used, where the danger mentioned cannot occur. Tungsten cathodes, however, require a much higher temperature for the same emission and a much higher cathode power than oxide cathodes. In order to supply momentary peak currents of 1—1.5 A, 125—150 W are needed with a tungsten cathode, compared with only about 8 W with an oxide cathode. In this case in valves of the type of fig. 1 temperatures of the tungsten wire of more than 2350 °C are reached. Since the evaporation of the tungsten is already very appreciable in this region of temperature (the vapour pressure lies in the neighbourhood of 10^{-6} mm of mercury) and as a consequence the filament has a lifetime of only a few hundred hours, the filament is often allowed to burn at a temperature of more than 100° lower during fluoroscopy (continuous operation), which already gives a lifetime of a couple of thousand hours, and the heating voltage is only increased to the necessary value for a moment just before each X-ray photographic exposure. Furthermore it is clear that fluctuations in the mains voltage have an unfavourable effect on the lifetime of the filaments when continually working so close to the limits. In order to improve this situation it is necessary to have recourse to the use of a stabilizer for the heating voltage.

The construction sketched in fig. 1 also presents difficulties as far as the reproducibility of the voltage loss is concerned. Secondary electrons from the anode strike the glass walls of the envelope, which are thereby charged and begin to exert a grid action on the current of electrons between cathode and anode. The potential to which the wall becomes charged depends upon all kinds of factors, for example on the occurrence of corona phenomena

outside the valve and the like. Although the relation between current transmitted and voltage loss of the valve at low measuring voltages is then often sufficiently reproducible, this is found to be no longer the case for operation under high-voltage. The voltage loss may then vary considerably and may sometimes be so high that the valve itself begins to emit X-rays to an appreciable extent and the anode becomes very hot.

Thus it is obvious that although the simple construction sketched might often answer well in installations for testing materials or for medical therapy and also for simple apparatus for diagnostics, where it is a question of maximum currents of a few hundred mA, it can certainly not be employed in every case in modern apparatus for diagnostics.

Gas-filled rectifier valves

An entirely different type of valve is found to satisfy better the requirements made of diagnostic apparatus. These valves are not highly evacuated but are filled with a gas to such a pressure that after ignition of the valve an arc discharge occurs. The rectifying action in this case is due to the fact that the ignition is initiated by the electron current emitted by the filament: in the half cycle in which the filament becomes positive normally there are not sufficient electrons and the valve does not ignite [3]).

Compared with vacuum valves we may conceive the situation during the arc discharge such that the space charge of the electrons emitted by the filament is compensated by positive ions of gas. Due to this, only extremely low voltages are necessary for passing even very large currents, for example 25 V at peak currents of 1.5 A. Such a small voltage loss can be entirely neglected with respect to the voltage on the X-ray tube.

The working voltage of the valve mentioned, in contrast to the case of the vacuum valves, depends little on the distance between the electrodes. Therefore, in constructing the valve there is no objection to have a rather large distance between the electrodes, which is of course desirable in order to avoid trouble from cold emission in the negative phase and to be able to use an oxide cathode.

Incidentally, in the case of gas-filled valves the cold emission is not the main cause of back-lash. Back-lash is more apt to occur due to a breakdown in the gas, resulting in an arc discharge in the wrong direction. The structural measures to be taken in order to avoid breakdown can be deduced from a consideration of the familiar Paschen curve, see *fig. 2*, which indicates the breakdown voltage V_d as a function of the product of gas pressure p and electrode distance d. If a given combination of gas

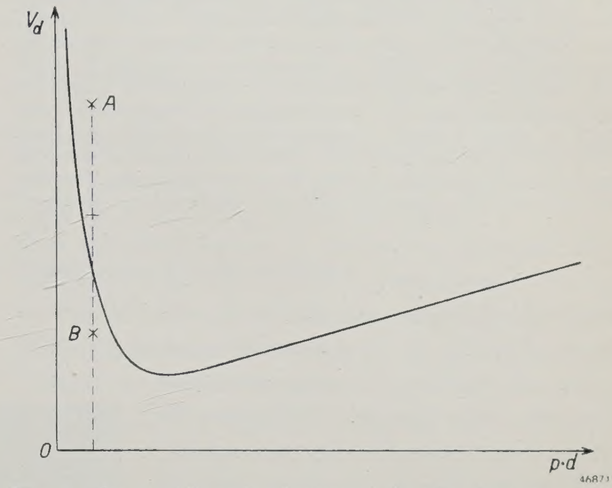


Fig. 2. Shape of Paschen's curve indicating for a given gas the breakdown voltage V_d as a function of $p \cdot d$ (p = gas pressure, d = distance between electrodes), for which, strictly speaking, two plane parallel electrodes are assumed. (Compare also Philips techn. Rev. **2**, 123, 1937 where it may be seen from fig. 2 that for argon, for instance, the minimum breakdown voltage of more than 250 V is reached at a value of more than 1 mm of mercury times cm for $p \cdot d$.) The rise in V_d at very small values of $p \cdot d$ is due to the fact that at those values the free path of the electrons becomes greater than the distance between the electrodes, so that the electrons can pass through the potential difference without causing any appreciable ionization. The line AB, the significance of which is explained in the text, actually passes much closer to the ordinate than it is drawn here, and the point B lies in practice at about a hundred times the voltage of the minimum of the curve.

pressure, electrode distance and valve voltage V (highest voltage between the electrodes occurring in the negative phase) corresponds to a point A, which lies above the curve, breakdown will occur. In order to avoid breakdown, when p and V are given the electrode distance should obviously be chosen much larger or much smaller. In practice the first method is out of the question, since then one would arrive at enormous lengths of the tube for the high-voltages required. But the second method brings us into conflict with the requirement that the electrode distance should be large enough to exclude cold emission.

A solution of this dilemna is found in a simple way. Suppose that at point A (fig. 2) cold emission were precluded. The valve is then built up of a number of units in series, each with the same electrode distance as for point A, and to each of these compartments only a proportional part of the total voltage is applied [4]).

[3]) For the general principles of gas-filled rectifiers see: M. J. Druyvesteyn and J. G. W. Mulder, Philips Techn. Rev. **2**, 122, 1937.

[4]) *Cf.* also J. G. W. Mulder, thesis Delft 1934.

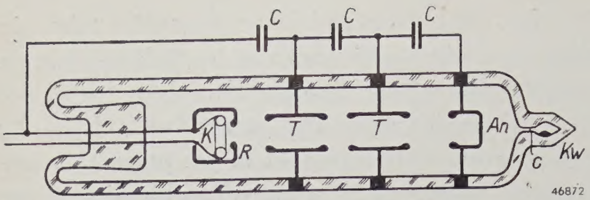


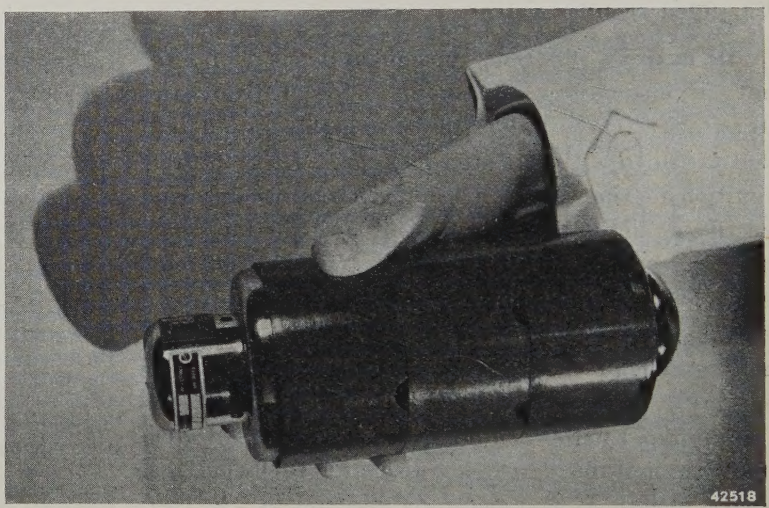
Fig. 3. Diagram of a cross-section of a gas-filled rectifier valve. The glass envelope is filled with saturated mercury vapour. The distance between the oxide cathode K and the anode An is divided into three stages by intermediate electrodes T, which are connected by the condensers C for the sake of a uniform voltage distribution. By a ring R placed in front of it the cathode is screened against bombardment by positive gas ions, while the ring at the same time lessens the chance of material from the oxide cathode striking the other electrodes. Kw is a mercury reservoir connected with the envelope by a capillary.

In this way one arrives at a work point for each separate compartment which, if the subdivision has been carried far enough, lies below the Paschen curve (B in fig. 2). Each compartment is then safe as far as breakdown is concerned, and since the field strength V/d (when the voltage is evenly distributed and the field homogeneous) is even smaller for point B than for point A^1, there is also no danger of cold emission. In practice, for example, a division of the valve into three stages is already sufficient. In *fig. 3* such a three-stage gas-filled valve is shown diagrammatically, while *fig. 4* shows two photographs of it.

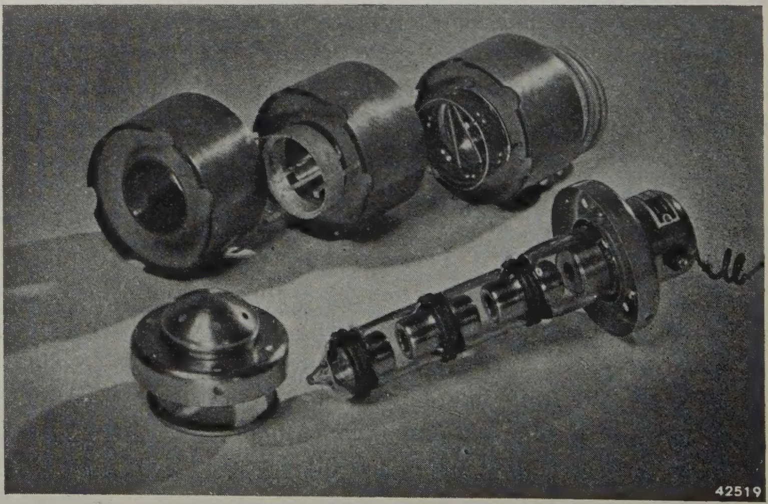
We have just spoken of the assumption of a homogeneous field and a uniform distribution of the voltage over the electrodes. In the construction according to fig. 3 the field is not, of course, entirely homogeneous; the intermediate electrodes cannot be constructed as parallel plane plates but should have holes in order to allow the passage of the electrons from the cathode to the anode. In practice they are constructed as cylinders. The greatest field strength now depends also on the shape of these cylinders and it is possible to influence the chance of back-lash, for instance, by the choice of the size of the opening of the cylinders. In order to guarantee a sufficiently uniform distribution of the voltage over the different stages the impedance between all successive intermediate electrodes should be made as nearly equal as possible. To that end condensers of 80 μF are connected in parallel with the intermediate spaces, which condensers have the form of rings lying around the rectifying valve, as may be seen in the photograph of fig. 4*b*.

Let us now turn to the gas with which the valve is filled. A rare gas cannot be used because of the well-known phenomenon of gradual disappearance of the gas during operation of the valve at high voltage: it is taken up in the cathode and walls; the pressure in the valve falls [5]). Therefore the discharge is made to take place in saturated mercury vapour: a drop of mercury is placed in the valve and from this the mercury vapour which disappears is always supplemented by evaporation. The result, however, is that the vapour pressure in the valve depends upon the temperature of the surroundings,

[5]) The velocity of this process (called clean-up) increases rapidly with increasing voltage. It is not the same for different gases; for xenon, for instance, it is smaller than for argon.



a



b

Fig. 4. A Philips three-stage gas-filled rectifier valve for 125 kV; *a*) assembled and *b*) taken apart. The glass tube is surrounded by ring-shaped "Philite" condensers (see C in fig. 3). The tube is calculated for oil immersion of the rectifier. The length is more than 20 cm, the diameter of the condensers is almost 9 cm. When used in air, in order to avoid flashover along the outside of the glass wall, a greater insulation distance between the electrodes would be necessary, thus in general a longer length of tube.

see *fig. 5*. Gas-filled valves can therefore only be used in a limited temperature range, which in the case of the Philips high-voltage rectifiers for X-ray purposes lies between about 17 and 40 °C. At too low a temperature the pressure becomes so low that no

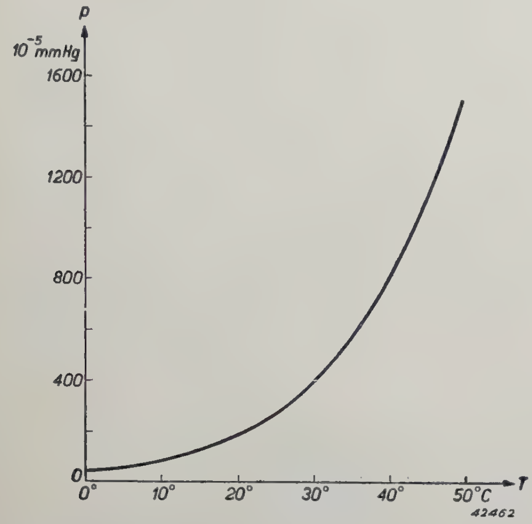


Fig. 5. Vapour pressure of mercury as a function of the temperature.

ignition occurs in the direction of transmission (the chance of formation of gaseous ions is too small); at too high a temperature the pressure becomes so high that breakdown occurs in the negative phase (the valve no longer rectifies). The latter can be seen directly from fig. 2: with increasing pressure the work point B is displaced horizontally to the right and after a certain pressure has been reached it comes to lie above the breakdown curve (unless B lies lower than the minimum of the curve, but in practice such a far-reaching subdivision of the voltage is not possible). The liquid mercury that has to be present in the valve should not come into contact with the anode or the intermediate electrodes, since local increases of the field strength would then occur and these might cause cold emission and thereby back-lash. Therefore the drop of mercury is placed in a special compartment, separated from the rest of the valve by a capillary (see fig. 3). Due to its surface tension the drop cannot flow through the capillary, while the mercury vapour is admitted freely to the valve.

Although, according to the above, we arrive at a distance between neighbouring electrodes in the gas-filled valves which is not greater than that in vacuum valves, it is, nevertheless, possible to employ an o x i d e c a t h o d e in gas-filled valves. Since there is no trouble with space charge here the cathode can be mounted behind a screen (the ring R in fig. 3) which prevents any particles shot off the cathode from reaching the neighbouring electrode. At the same time the screen prevents an ion bombardment of the hot cathode and the accompanying sputtering. As already stated, an oxide cathode requires a cathode power of only 8 W for the largest peak currents occurring; thereby the cathode temperature is about 900 °C. At this temperature it is found that even after some time no disturbing evaporation of barium from the oxide cathode occurs, although, remarkably enough, the vapour pressure of pure barium is quite considerable at the temperature mentioned.

Improved vacuum rectifiers

Let us return for a moment to the vacuum valves. The construction according to fig. 1 was not practicable for modern diagnostic apparatus. It proved, however, to be capable of improvement in various respects. An important improvement was the alteration of the construction as shown in *fig. 6*.

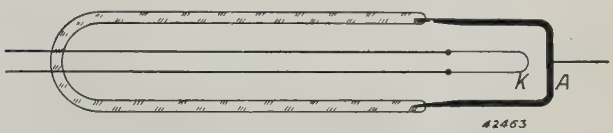


Fig. 6. The characteristic of the vacuum rectifiers can be considerably improved by the use of a cup-shaped anode A. In this case it forms part of the valve wall ("Metalix" rectifiers) which makes cooling easier.

Here the anode is not a flat plate at some distance from the cathode, but a cup surrounding the cathode. By this means the electrons having to pass from the filament to the anode are spread over a larger solid angle, the current density becomes smaller and therefore the space charge has less effect. In *fig. 7* the characteristic of a valve with

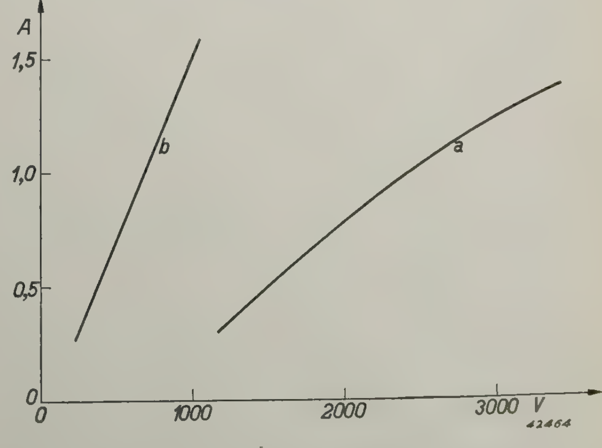


Fig. 7. Characteristics of vacuum rectifiers at high temperatures of the filament (more than 2350 °C): a) according to the old construction of fig. 1, b) according to the improved construction of fig. 6.

cup-shaped anode is compared with that of a valve with a plane anode. In order to transmit a current of 1 A the improved valve needs a voltage of only 650 V, compared with 2500 V in the case of the old valve. At the same time the cup shape of the anode, which in this case forms part of the wall of the valve, practically entirely prevents secondary electrons from striking the glass wall and there is no longer any fear of a grid action of varying wall charges.

A practical model of such a valve, the "Metalix" rectifier, to be used in air, is shown in *fig. 8*. These valves attracted considerable attention and they may still be found in use in existing diagnostic apparatus. Their further development, however, was curtailed by the introduction of the gas-filled valve. We have seen, indeed, that gas-filled valves are able to satisfy the highest requirements in diagnostics. The objection to gas-filled valves, namely the limitation of their usefulness to the narrow temperature range of 17-40 °C, makes gas-filled valves unsuitable for use in tropical climates or in apparatus for the macroscopic testing of materials, which have to be used under very divergent conditions of temperature. Thus in any case vacuum rectifiers, which are independent of temperature, still command a certain field of application, and for the sake of having a single type of rectifier of universal applicability further development of vacuum rectifiers, *viz.* a still better approximation of the requirements made by diagnostics, certainly seemed worthwhile.

In recent years several improvements have proved possible by which vacuum rectifiers, which had remained at the stage of development shown in fig. 6, could be given a new lease of life. Probably the most important improvement is the employment of thoriated tungsten for the filament. The addition of thorium to the tungsten results in a considerable increase in the electron emission, so that for the same tube current a lower filament temperature is sufficient and a smaller filament power is needed [6]). In practice a temperature of about 1750 °C is used. At this temperature the evaporation of the filament is so slight — although here again the vapour pressure of the pure thorium at the temperature in question is by no means low — that even after several thousand hours of operation no appreciable quantity of evaporated material can be detected as a deposit on the valve wall, not to speak of the occurrence of any burning through of the wire. The resulting, much lower, sensitivity to fluctuations of the heating voltage makes the above-described complications (increase of voltage for each exposure, stabilizer) unnecessary. The filament power is reduced from 125—150 W for the pure tungsten cathode to somewhat more than 30 W for the cathode of thoriated tungsten.

Fig. 8. "Metalix" vacuum rectifier valve. Since this valve is not intended for immersion in oil in the generator container, but for use in air, the envelope is made of non-transparent glass, in order not to give a disturbing light in the X-ray room. The anode is provided with cooling fins. The total length is 55 cm.

The thoriated tungsten filaments are especially sensitive with respect to the vacuum: traces of gas in the valve may reduce the electron emission considerably. Therefore the use of these cathodes is only possible by taking more care for a very good vacuum than in the case of the old vacuum valves. During manufacture the valves are very carefully evacuated. During operation the high vacuum is maintained with the help of a barium getter.

Another modification introduced in the construction of the "Metalix" rectifier valves of fig. 6, which made them more suitable for modern diagnostic apparatus, may be seen in *fig. 9*. The cup shape of the anode and thus the favourable characteristic

[6]) Because of the reduction in the filament power required, thoriated tungsten cathodes are also used in some modern transmitting valves. See E. G. Dorgelo, Several technical problems in the development of a new series of transmitting valves, Philips Techn. Rev. 6, 253, 1941.

of fig. 7*b* is retained, but now the anode no longer forms part of the valve wall. By means of a thin pin of material with a low heat conductivity (molybdenum) it is connected with a metal cap welded to a hard-glass envelope and serving for

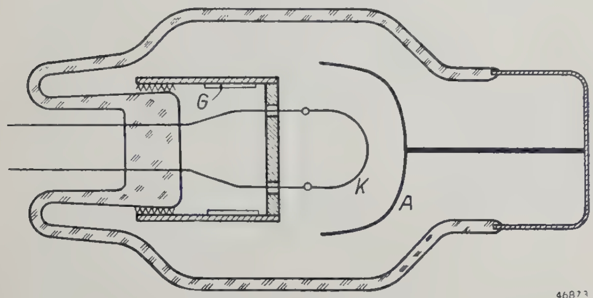


Fig. 9. Diagram of a modern vacuum rectifier. The filament *K* is made of thoriated tungsten. The cup-shaped anode *A* no longer forms a part of the (hard-glass) valve wall, which makes it easier to use the valve under oil. *G* is a getter (barium mirror).

the current supply. Thanks to this construction the anode can be heated to a higher temperature during degassing, while at the same time the field strengths along the glass are lower. By this means, and because of the fact that the metal cap in question remains relatively cold, it becomes easier — as in the case of the above-described gas-filled valves — to make the valves suitable for operation under oil. This leads to an important reduction in dimensions. The valve for 125 kV now has a length of somewhat more than 20 cm with a diameter of 9 cm (see *fig. 10*), *i.e.* about the same dimensions as the three-stage gas-filled valve.

Thanks to the cup shape of the anode in the new construction, practically no secondary electrons can strike the glass wall, and if this does happen the screening of the cathode is an adequate insurance against the occurrence of grid action. For the dissipation of the heat developed on the anode, which in the construction according to fig. 8 took place easily by conduction and convection into the surrounding air, in the new construction use is made of radiation. Because of the large radiating surface of the cup anode, an anode dissipation of 40 to 50 W, as occurs in continuous operation (about 30 mA), offers no difficulties at all, and a peak current of 1A, at which the anode dissipation amounts to about 600 W, can easily be coped with for a few seconds, since the anode possesses a sufficiently generous heat capacity.

Summarizing, we may say that vacuum rectifier valves in their latest forms give very interesting results and are already beginning to conquer considerable territory. It is possible that ultimately they will again entirely replace gas-filled rectifier valves.

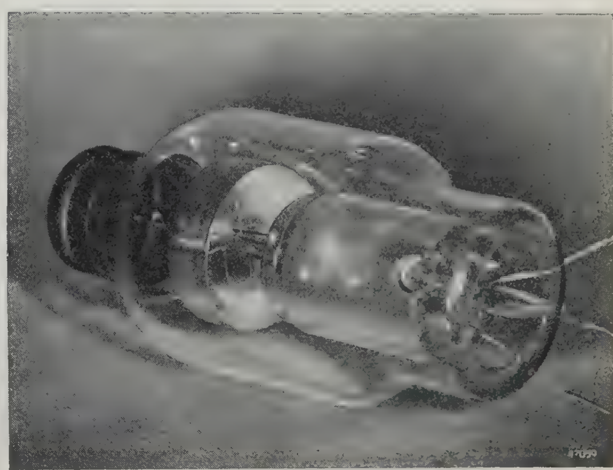


Fig. 10. Modern vacuum rectifier for 125 kV with cup-shaped anode and thoriated tungsten cathode. The valve, more than 20 cm long, is included to be used under oil.

CARRIER TELEGRAPHY

by J. te WINKEL.

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In order to improve the economy of telegraphic communication the aim used to be to increase the signalling speed by the construction of special apparatus or by employing devices for the transmission of several telegrams simultaneously over a single pair of conductors. The most modern form of the latter method is voice-frequency or carrier telegraphy, whereby alternating voltages of different frequencies (carriers) are each modulated in the rhythm of the signals of a telegram. After a discussion of the principles of this method an installation for carrier telegraphy is described in which 18 or 9 telegraphy channels, as the case may be, are contained in the frequency band of a single telephone connection, *i.e.* for example in a single channel of a carrier telephony system. Special attention is paid to the circuits employed for modulation of the telegraph signals on their carrier and those for their demodulation.

For many decades telegraphy, the oldest form of electrical communication, was the only available method for the transmission of messages over long distances. When later on, thanks to loading coils and repeaters, telephony could be used for covering large distances it took over a large part of the field of application of telegraphy. The advantage of being able to obtain an answer immediately, as well as that of personal contact, made for many purposes telephony more attractive than telegraphy. As a result the technique of electrical communication has recently been concentrated mainly on the perfecting of the telephone system, which has now attained a high standard of reliability and economy. Nevertheless, telegraphy has always retained its right of existence, especially in cases where it was desirable that the messages transmitted should be recorded in writing, and in general also because telegraphy is more economical and therefore cheaper to the public than telephony. It is remarkable that even the technical development of apparatus for telephony has now led to a new and especially economical telegraph system and thereby served to extend the field of telegraphy. This system, known as voice-frequency or carrier telegraphy, is based on the same principles and for the most part employs the same kind of apparatus as carrier telephony, which has been discussed in a series of articles in this periodical [1]).

Before proceeding to the description of a carrier telegraph system, such as has been developed and put into practical use by Philips, we shall give as an introduction a brief survey of the course of events in ordinary telegraphy, so-called direct-current telegraphy. This will give us the opportunity of mentioning and explaining several elementary concepts and quantities of telegraphy.

[1]) Philips Techn. Rev. **6**, 325, 1941; **7**, 83, 104 and 184, 1942, **8**, 137 and 168, 1946.

Direct-current telegraphy

Fig. 1 shows the circuit diagram of a direct current telegraph connection. At the transmitting station A by means of the switch S, for instance a signalling key, positive or negative voltages can be applied to the line. At the receiving station B the line is completed by a polar relay R. This relay reverses its armature as soon as the current through the relay winding changes its direction, the position of the contact arm T thus changing in the same way

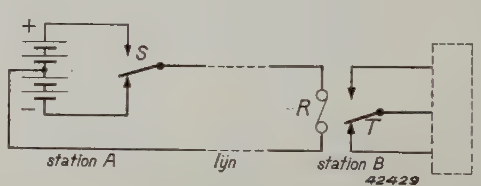

Fig. 1. Diagram of a connection between two stations A and B with direct-current telegraphy. S signalling key, R polar relay with contact arm T.

as the position of the switch S. *Via* the contact T some type of recording apparatus is operated from a source of voltage not shown in the diagram.

Signalling speed

For telegraphing over such a circuit a code is used: each letter is represented by a succession of a certain number of positive and negative voltage impulses of a certain length. The oldest and still commonly used code is the Morse code; for modern recording apparatus, for instance teleprinter apparatus, which are so constructed that each succession of impulses leads immediately to the printing of a certain letter, other codes are used. All the codes, however, have this in common, that the length of each impulse is chosen equal to a whole multiple of a fixed unit, the elementary length τ; see for example *fig. 2a*.

The speed at which one can telegraph is greater

the shorter the time devoted to the elementary signal length. The speed is expressed in baud units (from Baudot, one of the pioneers of telegraphy), *i.e.* the number of elementary lengths transmitted per second.

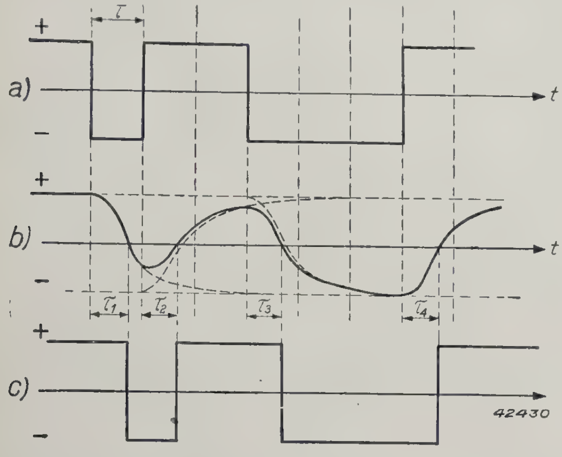

Fig. 2. Shape of the voltage curve at the transmitting end (a) and the current curve at the receiving end (b). At the zero points of the latter curve the voltage in the receiving apparatus is reversed in polarity by the relay contact T (c). Due to the fact that the "propagation times" $\tau_1 \dots \tau_0$ of the voltage changes differ more or less, c is not entirely identical with a: distortion has occurred.

The signalling speed is limited by properties of the line, which will be discussed later, and by the inertia of the transmitting and recording apparatus. For teleprinters, which are at present the most common, the signalling speed has been standardized by the C.C.I.T. (Comité Consultatif International Télégraphique) at 50 baud.

Signal distortion

The two conductors of the line possess a certain resistance and capacity which are distributed uniformly over the whole length of the connection. There exists also a certain self-induction and dissipation, which, however, can usually be disregarded. Every time the switch S in fig. 1 is reversed the line capacity is discharged and charged again with opposite polarity. The resistance of the line causes this charging and discharging to occupy some time; thus it will take some time after reversing the key at the transmitting end before the current through the relay at the receiving end reaches its final value. If we know the behaviour of this current for a single change of voltage, we can construct the curve of the current for each letter sign. In *fig. 2b* this has been done for the signal of fig. 2*a*. It may be seen that the current curve involves considerable distortion, namely a rounding off of all transitions. In actual practice it is not a question of the shape of the whole curve, but only of the position of the zero points. As soon as the current in the receiving relay passes through zero the contact T is reversed, and for good transmission of the signals it is sufficient when the contact T follows the key S in the correct rhythm. In *fig. 2c* the position of the contact T is shown as a function of the time for a relay current according to fig. 2*b*. If we compare this position with that of the switch S, it is found that each change of voltage possesses, as it were, a certain time of propagation (τ_1 to τ_4), the value of which is not the same for all changes. The result is that the length of the impulses which are fed to the recording apparatus is no longer exactly equal to the length of the impulses transmitted; the signals are distorted. Signal distortion is defined as the increase or decrease in the length of an impulse expressed as a percentage of the elementary signal length τ, thus

$$100\,\frac{\tau_2-\tau_1}{\tau},\ 100\,\frac{\tau_3-\tau_2}{\tau} \text{ and } 100\,\frac{\tau_1-\tau_3}{\tau}\ \%.$$

It is clear that signal distortion can also occur due to other causes, for example inequality of the positive and negative voltages, incorrect adjustment of the relay or accidental disturbances. The maximum distortion occurring is a measure of the quality of the connection. It may not exceed a certain value, for instance 10 to 20 percent, depending on the type of telegraph apparatus and the margin of safety chosen, so as not to incur the danger that impulses will be omitted and incorrect letters printed.

Economy of the connection and maximum distance to be covered

In order to use a given telegraph line as economically as possible it is desirable to raise the signalling speed as high as possible. For this purpose various systems of high-speed telegraphy have been designed and constructed in the past. From the above, however, it is seen that with the signalling speed ($1/\tau$) the signal distortion will increase. Furthermore the signal distortion (differences in the times of propagation $\tau_1, \tau_2, \dots$) will be greater, the greater the average time of propagation of the voltage changes. This time is determined by the RC time of the cable, and since R and C are both proportional to the length of the cable it increases proportional to the square of the distance to be covered.

It may therefore be seen that with a given limit for the permissible signal distortion, which limit one can try to place as high as possible by special circuits and constructions, a high signalling speed

leads to a limitation of the distance to be covered and, conversely, a given distance leads to a limitation of the signalling speed.

A different way of using a cable more economically was opened by systems which permitted several telegrams to be sent over the line simultaneously, each at a normal signalling speed. The most modern and at present the most important of these systems is the so-called voice-frequency or carrier telegraphy, which we shall now consider. It must, however, be noted at once, that here, too, is a limitation of the total signalling speed, a limitation which upon closer consideration proves to be based upon the same fundamental principles as in the case of direct-current telegraphy.

Voice-frequency telegraphy

In voice-frequency telegraphy the direct-current impulses are replaced by impulses of an alternating voltage of several hundred or thousand cycles per second. *Fig. 3* shows a simple circuit with which

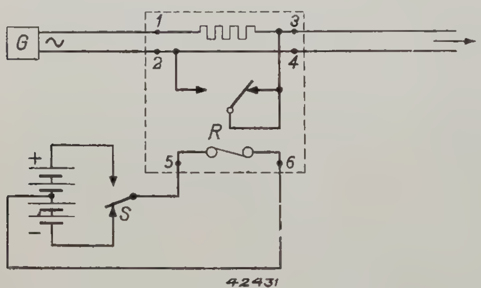

Fig. 3. Simple circuits for passing over from D.C. telegraphy to carrier telegraphy. C_1 generator of an A.C. voltage of several hundred or thousand c/sec, S signalling key, R polar relay which short-circuits or passes the A.C. voltage to be applied to the line in the rhythm of the signals given with S.

signals can be transmitted in this way, while *fig. 4* shows the appearance of a certain signal code in direct-current and in voice-frequency telegraphy. We may consider the alternating voltage as a carrier wave modulated in each case with the telegraph signals to be transmitted according to *fig. 4a* (the "depth of modulation" here is 100 percent). By using several carrier waves with different

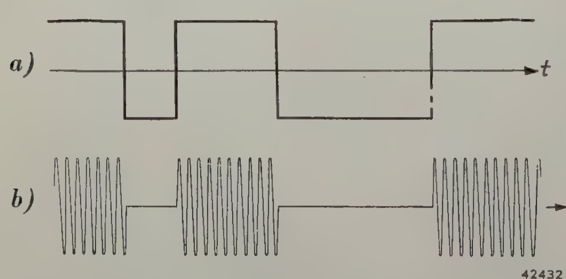

Fig. 4. The same signal code (succession of impulses) in D.C. and in carrier telegraphy.

frequencies, each of which is modulated separately with different signals, a number of telegrams can be sent simultaneously over the same pair of conductors. At the receiving end a series of selective receiving apparatus has to be provided each of which reacts only to one of the carriers.

Frequency band of a telegraph channel

Let us consider a signal in direct-current telegraphy which consists of alternating positive and negative impulses of the unit length τ, see *fig. 5*.

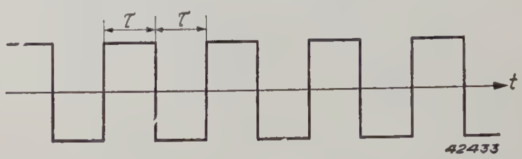

Fig. 5. Succession of positive and negative impulses with the unit length τ.

Such a signal may be conceived as an alternating voltage composed of a series of harmonics. At a signalling speed of 50 baud the fundamental frequency of this alternating voltage is 25 c/sec while further a large number (theroretically an infinite number) of higher harmonics occur. We may likewise assign to any given signal other than that of fig. 5 a complete frequency spectrum, beginning in the most general case with the frequency zero.

The distortion incurred by such a signal in being propagated over a telegraph cable can now also be described in a different way from that given in the foregoing. Due to its capacity and resistance the pair of conductors possesses an attenuation for the currents transmitted, which depends upon the frequency and increases with increasing frequency, since the capacity of the conductors forms a shunt whose impedance decreases with increasing frequency: the cable may be compared to a low-pass filter. In *fig. 6* the curve (attenuation loss *versus* frequency characteristic) of a normal telegraph cable is shown. A signal composed of a number of frequencies is now not only attenuated as a whole, but it is, moreover, distorted due to the fact that the higher harmonics are attenuated more than the

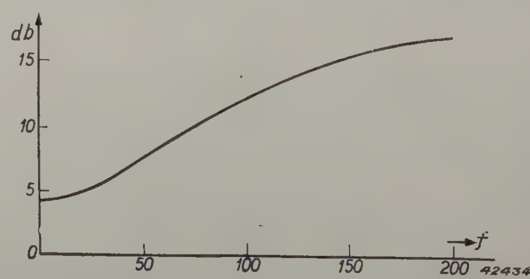

Fig. 6. Attenuation (in db) of a telegraph cable as a function of the frequency f in c/sec.

lower harmonics and the fundamental frequency. The effect of this prejudice to the higher frequencies is a rounding off of jumps and angles in the current curve, as we have seen in figs. 2*a* and *b*.

The relation between distortion, cable length and signalling speed can also easily be explained qualitatively in this way.

Let us now apply these considerations to voice-frequency telegraphy. When a signal with frequency q is modulated on a carrier with frequency p the two side-band frequencies $p \pm q$ occur. Thus in the modulation of a carrier by the telegraph signals two side-bands are formed, one on either side of the carrier frequency, which bands theoretically extend to infinity. These side-bands, however, need not be transmitted in their entirety: all the frequencies farther than a certain distance q_m from the carrier may be suppressed, for instance by means of a band-pass filter. This resolves itself in fact to the suppression of the higher harmonics in the signal, analogous to the unfavourable effect on those harmonics exerted by the cable in direct-current telegraphy. The result is again a distortion (rounding off) of the impulses transmitted, so that the modulated carrier assumes for instance the appearance of *fig.* 7. The degree of distortion per-

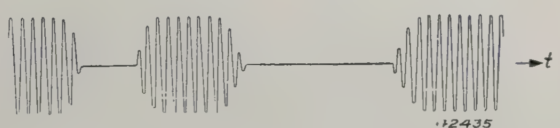

Fig. 7. The same signal code as in fig. 4*b* distorted as a consequence of the cutting off of the harmonics above 40 c/sec by the band-pass filter.

missible for reliable functioning of the receiving apparatus determines the minimum width ($2\,q_m$) which the transmitted frequency band must possess. It has been found that for a signalling speed of 50 baud in general a band width of $2\,q_m = 80$ c/sec is sufficient.

Several telegraph channels on one pair of conductors

If we wish to send several telegrams on different carriers simultaneously over a line without mutual interference we must assign to the carriers frequencies which lie sufficiently far apart, namely at least 80 c/sec, and use for each carrier a band pass filter with the correct transmitting band 80 c/sec in width. At the receiving end the respective carriers with their side-bands are split up again by means of a series of band-pass filters in parallel, identical with the transmitting filters. Following each receiving filter circuits are provided for the demodulation of the carrier, as well as a recording apparatus.

How many of these telegraph channels can now be included in a given frequency band of width Δf? If the loss of the band-pass filters at the boundaries of their transmitting region could jump abruptly to a sufficiently high value (at least 30 db), the distance between two adjacent carriers would not need to amount to more than 80 c/sec. Actually a filter always has a certain transition region between transmitting and attenuating regions, which transition can only be made narrower by a more expensive construction of the filter. In practice for example band-pass filters with the attenuation curve reproduced in *fig. 8* can be used. The neces-

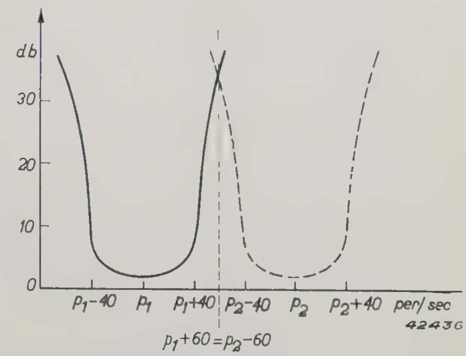

Fig. 8. Attenuation curve of a band-pass filter for the telegraph channel with carrier p_1. The broken line gives the corresponding curve for the adjacent channel with carrier p_1.

sary attenuation is here attained at a distance of 60 c/sec from the carrier, so that a carrier interval of 120 c/sec is required.

A comparison with carrier telephony as previously described shows that in that case — where the width of a channel is much greater (4000 c/sec) — the carrier itself and one of the sidebands are suppressed at the transmitting end. Due to the finite width of the transition region of the filters used, it is practically inevitable that the low frequency part of the transmitted side-band will also be cut off. In telephony this is no objection since the speech frequencies below 300 c/sec are not necessary for an intelligible conversation. In telegraphy, however, where all the modulation frequencies from zero onward have to be transmitted, it is not possible in practice to filter out only one side-band. For this reason the carrier and both side-bands are transmitted.

In a frequency band of width Δf, therefore, $\Delta f/120$ telegraph channels can be laid, so that one may speak of a "total" signalling speed of 50 $\Delta f/120$ baud. An obvious question is whether further profit can be gained by making the signalling speed in each channel greater than 50 baud. In the case of any given telegraph signal, however, all the frequencies of the composite alternating voltages are

proportional to the signalling speed, thus the required frequency band for a channel also exceeds 80 c/sec, and this proportionnally to the signalling speed. If band filters with an attenuation curve as in fig. 8 are always used, where only the frequency scale changes, so that the ratio between the width of transition region and transmission region remains the same, the total signalling speed is independent of the signalling speed in each channel. In other words, the total amount of "information" which can be transmitted is constant. It is of course true that with a greater band width of the filters the transition region can cable attenuation [2]) and amplification (or relaying) of the signals becomes necessary, which considerably increase the cost with increasing width of the frequency region. The situation here is analogous to that in carrier telephony, only in that case these considerations weigh even more heavily owing to the much higher modulation frequencies used in telephony.

Carrier telegraphy in a telephone channel

It is an obvious and much used method in cable telegraphy to make the total frequency band equal to that of an ordinary telephone connection (300-

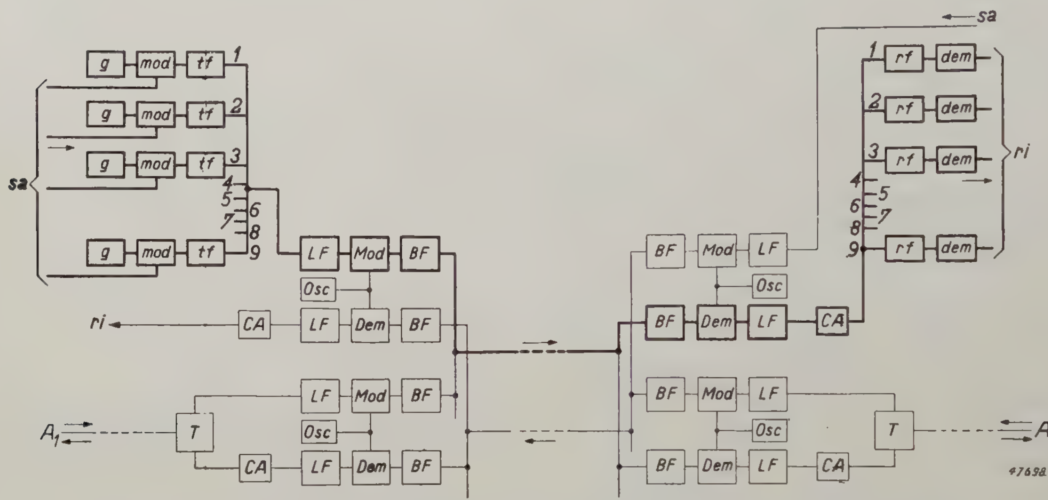

Fig. 9. Block diagram of a carrier telegraph connection using one channel of an installation for carrier telephony. Two channels of the latter are drawn, one for the carrier telegraphy (4 channels of which in one direction are indicated with heavy lines), the second for a normal telephone connection between two subscribers A_1 and A_2. The small letters refer to the telegraphy apparatus: *s.a* signalling apparatus, *r.i.* recording instruments, *g* generators, *mod* modulators, *tf* transmitting filters, *rf* receiving filters, *dem.* demodulators. The capital letters denote the components of the carrier telephone installation: *T* four-wire terminals, *LF* low-pass filters, *Mod* modulators, *BF* filters, *Dem* demodulators, *CA* channel amplifiers, *Osc* oscillators for the addition of the telephone carrier.

be made relatively narrower without much trouble, so that the signalling speed could be increased somewhat more. This would, however, by no means compensate for the complications and increase in cost which would accompany a higher speed of the transmitting and recording apparatus.

The attainment of a given total signalling speed therefore implies the use of a frequency band of a certain width, and how far one may go with this depends upon the circumstances. For example, if one telegraphs *via* a radio connection, the available frequency region is given by the band width of the radio transmitter. In the case of telegraphy *via* cables also an economic limit is set to the frequency region, since with frequency regions of any great size and with long distances, equalization of the 2600 c/sec). In this region, with a band width of 120 c/sec (and beginning at 360 c/sec, see below), no fewer than 18 telegraphy channels can be included.

These 18 telegraphy channels together can be transmitted in exactly the same way as a normal telephone conversation.

An especially economical telegraph connection is obtained when it is possible to reserve for carrier telegraphy one telephone channel of a carrier telephone installation with for instance 12 or more channels. The group of 18 telegraph carriers, each modulated with its own telegraph signals, is then modulated as a whole on one of the carriers of the telephone apparatus, as is indicated in the block diagram of *fig. 9*.

[2]) Philips Techn. Rev. **6**, and **7**, 293, 1941 184, 1942.

Since the large number of 18 telegraph channels will often be unnecessary, or since perhaps two instead of one of the many telephone channels in a given connection can be made available for telegraphy, Philips several years ago designed an apparatus whereby only 9 instead of 18 telegraph channels lie in the frequency band in question. This gives the advantage that the carrier interval may now amount to 240 c/sec, so that the band pass filters may have much wider transition regions and may therefore be simpler and cheaper.

As carrier frequencies in the frequency region mentioned from 300 to 2600 c/sec the values recommended by the C.C.I.T. may be taken: 420, 540, 660, etc. to 2460 c/sec, *i.e.* odd multiples of 60 c/sec. The odd multiples have the advantage over the even multiples that second harmonics and sum and difference frequencies of the carriers, which frequencies occur due to non-linear distortion in amplifiers etc., fall just between the carriers where the attenuation of the band-pass filters is high.

Apparatus for carrier telegraphy

The apparatus for carrier telegraphy, as we have seen, contains for each channel a carrier generator, a modulator and a filter for transmission, as well as a filter and a demodulator for reception. A series of other elements is also needed, such as relays, regulators, meters for adjusting the currents, etc. We shall consider here only two components of the apparatus in somewhat more detail, namely the modulator and the demodulator.

Modulator

For the modulation of the telegraph signals on the carrier, instead of the relay used in fig. 3 a device can be used which contains no moving parts and thus has the advantage that neither maintenance nor adjustment is necessary. This device, the telegraph modulator, is practically identical with the modulators used in carrier telephony which have already been discussed in this periodical [3]). The diagram illustrating the principle of the telegraph modulator ("blocker") was also described there, and the circuit diagram is given once more in *fig. 10a*. To the terminals *1-2* the voice-frequency alternating voltage is applied which serves as carrier, to the terminals *5-6* the direct current impulses of the telegraph signals. The resistance of the selenium rectifiers s_1, s_2 for the alternating current depends upon the polarity of the direct voltage acting on them at the same time. If point *6* is positive, the resistance in question is high; if point *5* is positive, it is low. The alternating voltage at the output terminals *3-4* will thus alternate in intensity according to the rhythm in which the voltage on *5-6* is reversed in polarity.

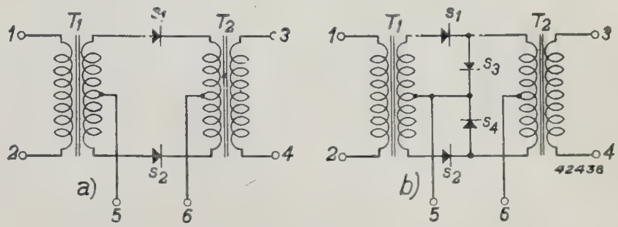

Fig. 10. Modulator connections. T_1, T_2 transformers, $s_1 \ldots s_4$ selenium rectifiers. At *1-2* the carrier is applied, at *5-6* the D.C. telegraphy signals to be transmitted, while at *3-4* the modulated signal is taken off. The modulator replaces in fig. 3 the circuit situated inside the block within the broken line, with the correspondingly numbered connections *1* to *6* inclusive.

Fig. 10b shows a somewhat modified circuit which, as may easily be seen, has the property that the resistance of the direct-current path between the points *5-6* upon reversing the polarity remains practically constant; this is desirable because the signalling apparatus to be connected at *5-6* is in general constructed for a specific external resistance.

In carrier telephony the audio-frequency microphone currents are applied to the terminals *1-2* of fig. 10*a*, the high-frequency carrier to *5-6*. In our case, on the other hand, the (audio-frequency) "carrier" is applied to *1-2*, while the much more slowly changing D.C. signals are applied to *5-6*. The reason for this difference lies in the fact that in telephony the carrier itself need not be transmitted; this is accomplished by balanced connection with respect to points *5-6*. In telegraphy the carrier has to be transmitted; moreover, the low-frequency telegraph signals, if they were applied to *1-2*, would not be passed on by the transformer T_1.

Demodulator

At the receiving end of the telegraph system the telegraph carriers, after being split up by the receiving band-pass filters, must each be demodulated separately in order to reproduce the telegraph signals. Whereas for the demodulation in carrier telephone channels, containing only one side-band without carrier, the carrier itself must be added again [4]), for the demodulation of the telegraph channels this is not necessary. These channels contain the carrier and both side-bands, so that the demodulation resolves itself into a simple detection, as is employed for instance in radio receiving sets.

[3]) Philips Techn. Rev. 7, 83, 1942.

[4]) See for example Philips Techn. Rev. 8, 137, 1946.

A circuit suitable for this purpose, working with anode detection, is shown in *fig. 11a*. The amplifier valve B has such a large negative grid bias that in the absence of alternating voltage on the grid the anode current is exactly zero (class B adjustment). Now suppose that the telegraph signal shown in fig. 5 is being transmitted. It consists of positive and negative impulses of the unit length τ. An alternating voltage is applied to the grid as shown in *fig. 11b* (left, below), whereby the envelope of the signal, distorted in a certain way, is assumed for the sake of simplicity to consist of straight lines.

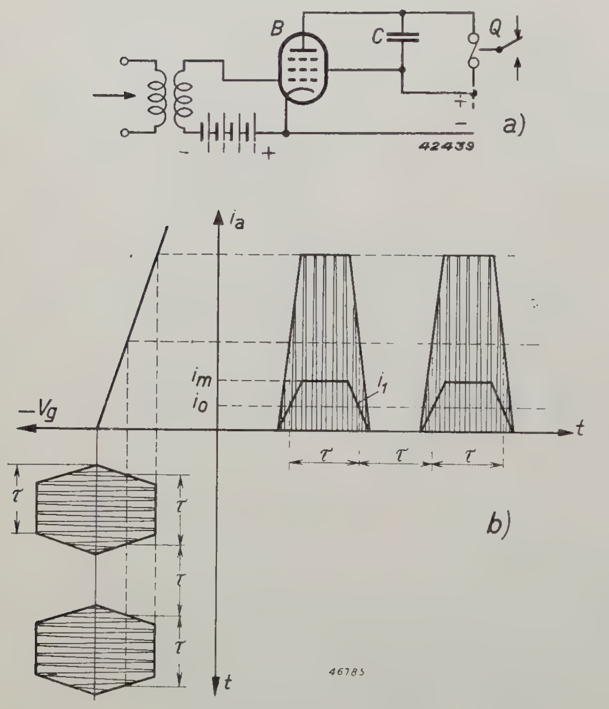

Fig. 11. *a*) Simple connections for the demodulation (detection) of the carrier telegraph signals. The valve B works in class B adjustment.
b) Form of the grid A.C. voltage V_g and the anode current i_a in the case of a signal consisting of impulses of the unit length τ. The low-frequency component i_1 of the anode current flows through the relay Q. This is so adjusted that it reverses its armature at a current $i_0 = i_m/2$.

If we also assume a straight line for the valve characteristic, the anode current i_a takes the form shown in the figure. It can be split up into a low-frequency component i_1, which flows through the receiving relay Q, and a component with carrier frequency, which flows through the condenser C, connected in parallel with the relay. The low-frequency component i_1 is of similar shape as that of the envelope of the anode current (namely divided by a factor π). It may be seen from the figure that the original telegraph signal is reproduced undistorted provided the relay Q is so adjusted that it reverses its armature at a relay current i_0 amounting to one half of the maximum value i_m. (In case the envelope should not be straight, a different but in any case very definite fraction would take its place.) If Q is a polar relay this adjustment can be realized by sending a constant direct current through an auxiliary winding of the relay in such a way that the field of the current i_0 is just cancelled. A simpler solution and one therefore actually employed in our apparatus is to take for Q a normal (non-polar) relay and to adjust the tension of the spring holding the armature in position to the correct value.

Compensation of changes in level

The detection circuit described only functions free of distortion as long as the level of the carrier received remains constant. Actually that level fluctuates more or less, due for example to the fact that the cable attenuation varies with the temperature. If the normal level is such that the low-frequency anode current i_1 flowing through the receiving relay has the shape $ABCD$ in *fig. 12*,

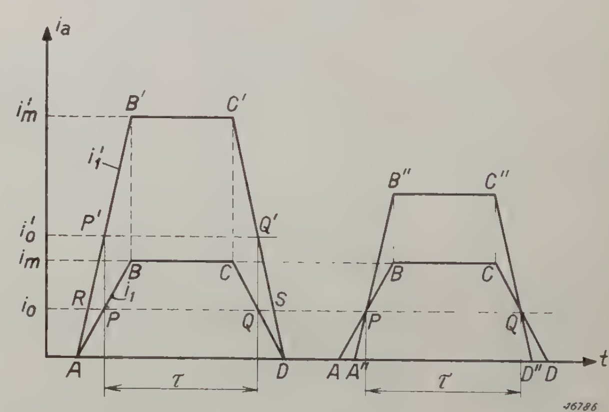

Fig. 12. If the low-frequency component i_1 of the anode current increases to i_1', the impulse received is too long, namely RS instead of PQ. This distortion can be compensated for by decreasing i_1' by a direct current $i_0'-i_0 = (i_m'-i_m)/2$.

at a higher level for example the shape AB^1C^1D will occur. The signal received is now too long, namely RS instead of PQ. A signal distortion thus occurs to the amount of $100\,\dfrac{RS-PQ}{PQ}\,\%$.

This distortion would be cancelled if the line AB^1C^1D could be lowered by the amount $PP^1 = {}^1/_2\,BB^1$. The shape of the current curve would then be $A''B''C''D''$, which intersects the original line $ABCD$ at the height i_0, thus at the correct points P and Q.

The signal-frequency anode current i_1' must thus be decreased by a direct current i_0'-i_0 equal to one half the increase of its maximum value (i_m'-i_m). A circuit diagram which accomplishes this is shown

schematically in *fig. 13a* [5]). In the grid circuit of the amplifier valve an extra rectifying circuit is introduced, consisting of the selenium rectifier s and the resistance R_1, shunted by the condenser C_1. The valve is connected between a tap *1* at the middle of the input transformer and a certain tap *3* on the grid voltage battery. In the non-signalling state (no grid A.C. voltage) a D.C. voltage V_2 therefore acts on the valve in the blocking direction. When a certain A.C. voltage (carrier) whose amplitude is smaller than 2 V_2 acts on the grid, no current will ever flow through the selenium rectifier, since point *1* never becomes positive with respect to point *3*. If, however, due to an increase in the receiving level, the amplitude of the grid A.C. voltage becomes greater than $2V_2$, namely $2\,(V_2 + \Delta V)$, a current flows through the rectifier at the positive peaks. The condenser C_1 is thereby charged, point *2* becomes negative with respect to point *4* and this proceeds until *1* and *3* again have the same potential at the peaks of the grid A.C. voltage. This is the case at a condenser voltage of ΔV. The grid bias therefore automatically becomes more negative by the amount ΔV, *i.e.* half the increase of the carrier amplitude; see *fig. 13b*. It may immediately be seen from the figure that also the envelope of the anode current and thus also the signal-frequency component of the anode current is lowered by half the amount with which its maximum value would have increased in the absence of s, C_1 and R_1. This is exactly the shift which was desired.

If therefore the voltage V_2 tapped from the battery is made equal to half the peak value of the carrier at the lowest level occurring, no signal distortion occurs at higher levels. In this way level variations to 6 db above and below the normal level can be permitted. The accuracy of the regulation is only limited by the time constant R_1C_1 which must be made large enough for the extra grid voltage furnished by C_1 to be retained in the intervals between the impulses, *i.e.* where the carrier amplitude is zero for a time. In the sometimes long intervals between successive telegrams, thanks to a suitable connection of the transmitter the carrier is transmitted continuously (in conformity with the recommendation by the C.C.I.T.), so that the grid voltage is set at the correct value immediately a telegram starts.

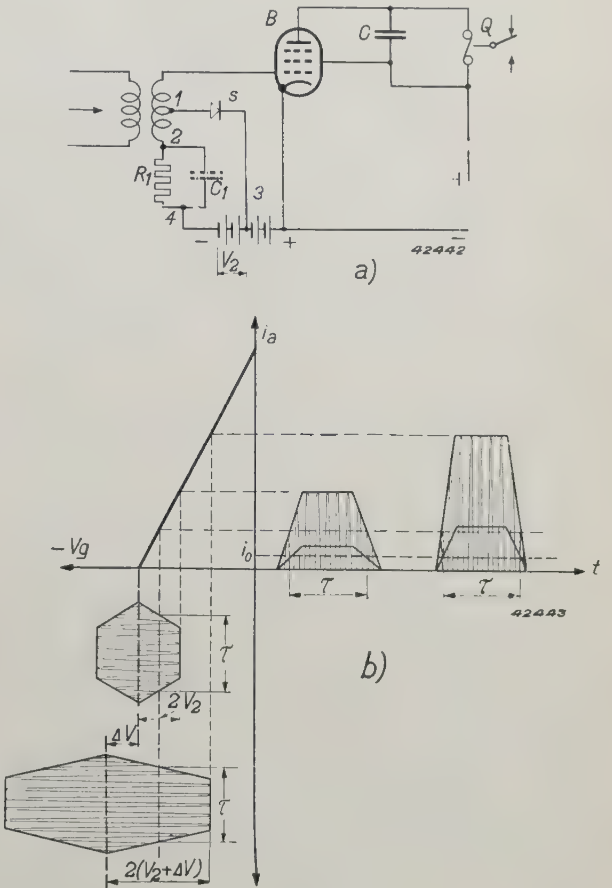

Fig. 13. *a*) Detection circuit with automatic compensation of the signal distortion caused by changes in level according to fig. 12; *s* is a selenium rectifier. Tap *3* of the grid bias is so chosen that $2V_2$ is equal to the amplitude of the grid A.C. voltage at the lowest receiving level occurring.

b) Explanation of the functioning of the circuit shown in the diagram. If the amplitude of the grid A.C. voltage rises from $2V_2$ to $2\,(V_2 + \Delta V)$, by the charging of the condenser C_1 the grid bias becomes more negative by the amount ΔV.

[5]) A similar compensation for changes in level is necessary in signalling in a telephone installation, see Philips Techn. Rev, **8**, 168, 1946. The solution there described is somewhat simpler than the one given here. The difference is due to the fact that the impulses received in telephone signalling (dialling impulses) have much steeper fronts than the impulses in telegraphy. With a steeper front the distortions occurring and the required corrections are obviously smaller (*cf.* fig. 12).

VELOCITY-MODULATION VALVES

by F. M. PENNING.

621.385.83

When an ordinary radio valve is used as oscillator, below a certain wave length the efficiency rapidly approaches zero. This behaviour is caused partly by the transit time of the electrons. If one attempts to shorten the transit time by reducing the valve dimensions, the energy which the valve can deliver again decreases rapidly. In order to avoid this difficulty in recent years a different principle has been applied for the excitation of decimeter and centimeter waves, whereby the transit time of the electrons is used to advantage. The beam of fast electrons concentrated by a magnetic or an electric field is made to pass through a longitudinal alternating field which produces a velocity modulation of the beam. As a result electrons which left the source later can overtake those which left earlier and fluctuations of density begin to appear along the beam. The beam is then passed through the openings of a cavity resonator and later received on a large anode. As a consequence of the fluctuations in density electromagnetic oscillations are caused in the cavity resonator by induction. By back-coupling these oscillations to the part where the modulating alternating field is excited — also a cavity resonator or a Lecher system — an oscillator can be realized. In this article it is explained how the electron concentrations depend upon the time and the place in the "velocity-modulator valve", and how with several simplifying assumptions the efficiency can be approximately calculated from two characteristic quantities: the ratio ϱ of external resistance to valve resistance and the product of the number ξ of the concentrations or parcels of electrons in the velocity-modulating space and the back-coupling factor K. For the sake of clarity an example is discussed.

When more than a half a century ago Heinrich Hertz carried out his classical experiments on electromagnetic waves he used a wave length of about 30 cm. For various reasons radio technology developed first mainly in the direction of very long waves; only in recent years has there been greater interest in radio waves of a few decimeters down to a few centimeters.

When ordinary radio valves are used for the excitation of very short waves one encounters various difficulties, as has often been explained in this periodical, difficulties which are caused for example by the fact that the transit times of the electrons are no longer negligibly small compared with the oscillation time, as at low frequencies [1]). This causes a damping in the grid circuit, with the result that oscillations of too high a frequency can no longer be amplified and the valve can no longer oscillate above a certain frequency. Attempts may be made to meet this difficulty by reducing the dimensions, but this is naturally limited, and moreover small dimensions involve limitation to small energies.

Attempts have been made to turn these transit time effects, detrimental in ordinary oscillators, to advantage by changing the construction of the valve or its connections.

This can be done in several ways. The oldest known example is found in the valves according to Barkhausen and Kurz; another example is the magnetron, which was discussed in a previous article in this periodical [2]). In both cases the electrons swing back and forth in the valve. The distances and velocities are so chosen that on an average as much energy as possible is taken per period from the electrons by the alternating field, which energy then appears as energy of oscillation.

Use is made of transit times in a different way in the so-called velocity-modulation valves [3]), which are the subject of this article. In these valves the electrons do not swing back and forth but always move in the same direction. The modulation of the electron current is here produced at the entrance to the tube by periodic variation of the velocity, previously equal for all electrons, as a function of the time. This results in periodic current variations farther along in the tube, because when the electrons which started later have a higher velocity they overtake those which started earlier.

Finally there are oscillator valves in which use is made simultaneously of the velocity modulating

[1]) See for these transit times, for example, C. J. Bakker, Some characteristics of receiving valves in short-wave reception, Philips Techn. Rev. 1, 171, 1936; M. J. O. Strutt and A. van der Ziel, The behaviour of amplifier valves at very high frequencies, Philips Techn. Rev. 3, 103, 1938.

[2]) Philips Techn. Rev. 4, 189, 1939.

[3]) The literature on velocity modulator valves has become very extensive in recent years. We mention here only the first articles by the pioneers in this field. W. C. Hahn and G. F. Metcalf, Proc. Inst. Radio Eng. 27, 106, 1939; R. H. Varian and S. Varian, J. Applied Phys. 10, 321, 1939; D. L. Webster, J. Applied Phys. 10, 501 and 864, 1939.

effect just described and of the swinging back and forth of the electrons, which is characteristic of Barkhausen-Kurz valves. These oscillator valves are also very important in practice. An interesting example will shortly be discussed in this periodical.

Before we now pass on to the elucidation of the principles on which the functioning of velocity-modulating valves is based, we shall make a few general remarks about the way in which the high-frequency energy of the electrons is transmitted to the oscillation circuit.

In ordinary transmitting valves the electrons finally strike an anode forming part of the oscillation circuit, which with the small dimensions of the valves results in the above-mentioned limitation of energy. On the other hand in the velocity-modulator valves and also in other oscillator valves based upon transit time effects, the electrons are made to pass along the "oscillation circuit" (a cavity resonator is often used for this) and give off their energy inductively [4]). The anode is placed farther on in the valve and may have any desired size.

Principle of the velocity-modulator valve

Fig. 1, which also gives the notation used, shows diagrammatically the arrangement of a velocity modulator valve. The oscillator system contains two grids A_1B_1 (modulator) placed very close together and considered to be absolutely permeable, and another similar pair A_2B_2 (inductor) separated

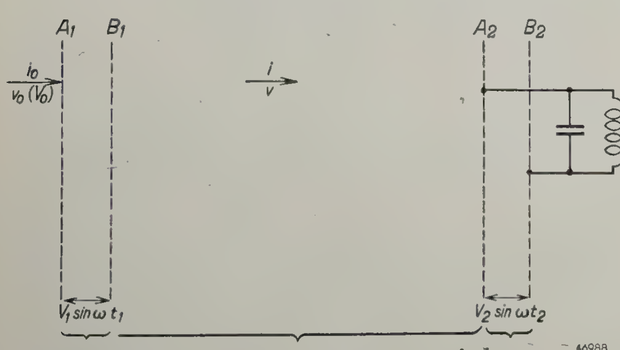

Fig. 1. Diagram showing the principle of the velocity modulator valve. A beam of electrons with a current i_0 amperes and an energy V_0 electron volts (velocity v_0) first passes two closely placed grids A_1B_1 (modulator), between which an A.C. voltage $V_1 \sin \omega t$ is applied.
The velocity modulation of the electrons thereby excited is converted in the "overtake space" B_1A_2, which is free of field, into a modulation of the convection current i. This modulated current gives off energy to an oscillation circuit (inductor) connected between the grids A_2 and B_2, which is represented diagrammatically by a self-induction and a capacity, although a separate L and C are not used, but for instance a cavity resonator.

[4]) See for example p. 159 of the article by C. G. A. von Lindern and G. de Vries "Flat cavities as electrical resonators" Philips Techn. Rev. 8, 149, 1946.

from A_1B_1 by a space free of field B_1A_2 (velocity-modulating space). The grid A_1 is at a potential V_0 with respect to the cathode. If an *A.C.* voltage $V_1 \sin \omega t_1 = aV_0 \sin \omega t_1$ is now applied between B_1 and A_1 ($a = V_1/V_0$ = depth of modulation), the voltage on B_1 is equal to $V_0\ (1 + a \sin \omega t_1)$. The electrons arrive on the grid A_1 with a velocity v_0, which corresponds to the voltage V_0, and leave grid B_1 with a velocity v, which depends upon t_1 according to the relation

$$v = v_0 \sqrt{1 + a \sin \omega t_1} = v_0 \sqrt{1 + a \sin 2\pi t_1/T} \quad (1)$$

(T = time of oscillation). In this way the beam of electrons receives a velocity modulation. For $-T/4 < t_1 < T/4$, v increases with t_1, and thus electrons leaving B_1 later will be able to overtake the electrons leaving B_1 earlier. This can be seen clearly in *fig. 2*. The vertical x axis represents the direction of motion of the electrons. The modulator A_1B_1 lies at the point $x = 0$. We assume that at regular intervals of 1/12 T electrons with equal velocity v_0 move from below towards the system A_1B_1. In the diagram of fig. 2 the distance x covered by each electron can be read off as a function of t.

The slope of the lines is here a measure of the velocity of the electrons. In the lower part of the figure it amounts to 45°, since the quantity x/v_0T is plotted along the vertical axis and along the horizontal axis the quantity t/T. The lines representing the behaviour of the different electrons are therefore parallel at the bottom and also equidistant. Due to the velocity modulation this is no longer true in the upper part of the figure. Upon passing the point $x = 0$ the lines in question will therefore show a bend the degree and direction of which are determined by the depth of modulation a and the moment t_1 at which the electron passes the modulator, that is the point $x = 0$. (For a given line t_1 is the abscissa of the point of intersection with the t-axis; t_1 is therefore the moment at which the electron that was at x at the moment t left the modulator ($x = 0$).) Due to the modulating voltage these bends will form a pattern in the part of the figure above the line $x = 0$ in which intersections of the lines in question occur, which means that the electrons overtake each other. The character of this pattern depends upon the value of the depth of modulation a. In fig. 2 the value $a = 0{\cdot}2$ has been chosen.

We are now able to follow the course of a given group of electrons in their passage through the valve. We draw for example a vertical line at $t = {}^1/_2\, T$ and cut it by the 45° line through the origin and 12 neighbouring lines. The ordinates of the

intersection points, given in the figure by dots, represent the positions of 13 equidistant electrons which have not yet reached the modulator at the moment $t = -{}^1/_2T$. In order to discover what happens to this group of electrons at later moments, $t = 0, {}^1/_2T, T$ etc. we must cut the bent lines by the vertical lines corresponding to these moments. These points are also indicated in the figure and it is zontal line the larger the current at the corresponding moment. From the figure it can be read off quite easily that for a given value of x the current is a purely periodical function of time, but that for different values of x the character of this periodical function differs.

The model of *fig. 3* shows us the form of this periodic function for a number of values of x/v_0T.

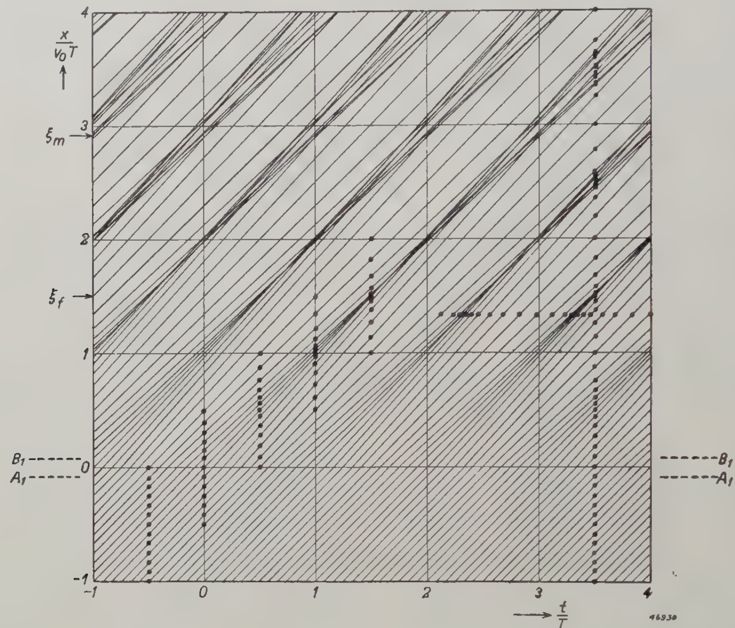

Fig. 2. Distance-time diagram for electrons which pass the grids A_1B_1 of fig. 1 at regular time intervals $T/12$. The quantities x/v_0T and t/T are plotted. Each of the continuous oblique lines represents the behaviour of x as a function of t for a given electron. The bend which the lines exhibit at $x = 0$ is determined by the depth of modulation a, for which in this case the value 0.2 is chosen. It may be seen how concentrations and rarefactions occur in the beam. At ξ_f (focus) the current first becomes infinite; at ξ_m (maximum plane) the first Fourier component of the A.C. is a maximum.

clear how a concentration gradually occurs in the middle of the group.

This concentration takes place around the electron which passes the modulator at the time $t = 0$, whose velocity v_0 thus remains unchanged. The concentration is thus likewise propagated through the valve at the velocity v_0. It increases in strength and at about $t = 1{}^1/_2\ T$ reaches a maximum, while for $t > {}^1/_2T$ the order of several electrons is found to have changed.

In a similar way it is possible to study an intersection of the network of lines of fig. 2 with a horizontal line, for instance the line $x/v_0T = 1$. This gives an idea as to what takes place at a given point in the valve in the course of time. The points of intersection of the chosen horizontal line with the oblique lines represent the moments at which the various electrons pass the point in question. It is clear that the closer the intersections lie on the hori-

(In accordance with the abbreviations to be introduced presently, the quantity x/v_0T is indicated in the figure by ξ and t/T by τ.) Fig. 3 shows more clearly than fig. 2 how a concentration, which, for example, begins to form at the moment $t = 0$ and at $x = 0$ is propagated through the valve in the direction of increasing x, increases in density. The distance-time line of the current concentration makes an angle of 45° with the horizontal axis, which means — in reference to the scales chosen — that the concentration moves through the valve with a velocity v_0, as already noted above.

Fig. 4 is a reproduction of an instantaneous picture of the current through the valve (cross-section of the surface of fig. 3 for $\tau = 1.52$, the moment at which the current first becomes infinite). It may be seen how the centres of the current concentrations are separated from each other by a distance of about $\lambda_1 = v_0T$, which may be called

the "concentration wave length". Since the phenomena are repeated periodically with the oscillation time T, the instantaneous pictures, differing only by a time T, are identical. The concentrations thus travel through the valve as a kind of wave with an

Fig. 3. The variation of the current i in a velocity modulator as a function of the time t and the distance x to the modulator. Plotted horizontally are the values of $x/v_0\ T$ ($=\xi$) and t/T ($=\tau$), vertically the value of i/i_0 where i_0 is the unmodulated current. The upper edges of the vertical strips thus indicate how the current at a given place varies as a function of the time. The current maxima are propagated at a velocity v_0 in the longitudinal direction of the valve (ξ direction). They meanwhile first increase in height and later split up into two peaks which move farther and farther apart.

oscillation time T, a velocity of propagation v_0 and a wave length λ_1. The height of the top of the wave thereby increases and becomes infinite at a certain spot. The peak then splits up into two parts which become farther and farther apart. It is to be noted that the electromagnetic oscillations of the same oscillation time T have a velocity c and a wave length $\lambda = cT$. For the relation between λ_1 and λ the following relation is apparently valid:

$$\lambda_i/\lambda = v_0/c = 0{,}002\sqrt{V_0}\ (V_0 \text{ in volts}) . \quad (2)$$

In the foregoing and with reference to figs. 3 and 4 we have discussed the behaviour of the electron current i as a function of the distance x from the modulator and of the time t. We shall now show how this function and thus also the figures can be calculated. We shall then particularize the calculations for the case, so important in practice, where the velocity-modulator valve oscillates at the fundamental frequency ω and the depth of modulation a is small ($a \ll 1$). The results of these calculations will enable us to discuss the theory of the velocity modulator as amplifier and as oscillator and to explain it with an empirical example.

Calculation of the current variations

For the calculation of i as a function of x and t it is helpful to use as auxiliary quantity the time t_1 already introduced. Obviously

$$x = v\ (t - t_1),$$

where v, the velocity of the electron, is given by equation (1). When this value of v is substituted in the above equation we find for the relation between t_1, t and x:

$$t = t_1 + x/v = t_1 + x/v_0 \sqrt{1 + a \sin \omega t_1} . \quad (3)$$

The latter equation and those following can be more simply written when one uses as variables, instead of the distance x and the time t, the non-dimensional quantities used already in figs. 2, 3 and 4.

$$\xi = x/v_0T \text{ and } \tau = t/T \ . \ . \ . \ . \ . \quad (4)$$

Then according to (2)

$$\xi = x/v_0T = x/\lambda_i = 500\ x/\lambda\sqrt{V_0}.\ (V_0 \text{ in volts}) \quad (5)$$

The quantity ξ has a simple physical significance. Since the distance between the centres of two current concentrations according to fig. 4 amounts to λ_1, for the case where ξ is a whole number $\xi = x/\lambda_i$ represents the number of concentrations or parcels situated within the distance x. With an obvious extension even when ξ is not a whole number it can still be called the "number of parcels" in the distance x; we then speak, for example, of $^3/_4$ parcel in the distance d when the distance between two parcels is $4/3\ d$.

With the help of the new variables ξ and τ equation (3) may be written as follows:

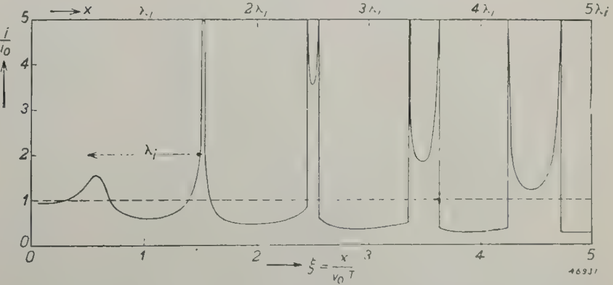

Fig. 4. Instantaneous picture (for $\tau = 1.52$) of the current variation in the valve as a function of the distance x to the modulator. The centres of the current concentrations have a mutual distance approximately equal to $\eta_1 = v_0T$ and are propagated in the valve as a wave of wave length λ_i, velocity v_0 and oscillation time T.

$$\tau = \tau_1 + \xi/\sqrt{1 + a \sin 2\pi\tau_1} \quad . \quad . \quad . \quad . \quad (6)$$

This relation between the reduced times of arrival (τ) and departure (τ_1) is shown in *fig. 5* for a number of values of the reduced distance ξ (it is assumed that $a = 0.2$). It may be seen in the figure that for sufficiently large values of ξ, namely $\xi > \xi_f$, three values of τ_1 correspond to one value of τ, *i.e.* at a single given moment, τ, electrons pass the spot in question ξ, which left the modulator at three different moments τ_1. On the other hand for $\xi > \xi_f$ only one value of τ_1 corresponds to one value of τ. We shall revert to the significance of ξ_f later on.

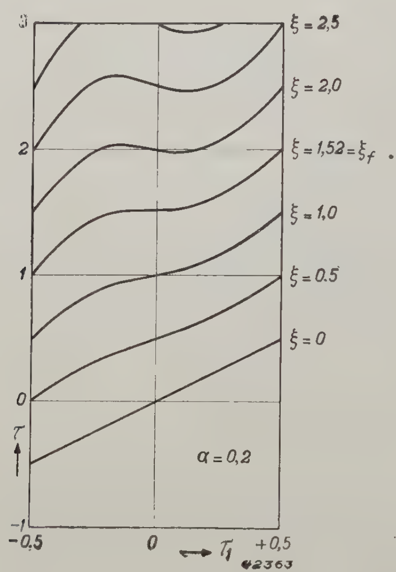

Fig. 5. Reduced time of arrival τ of electrons, which have covered a reduced distance ξ, as a function of the reduced time of departure τ_1. The scale for τ_1 is taken twice as large as that for τ. Depth of modulation $a = 0.2$. At $\xi = 1.52$ the value of $d\tau_1/d\tau = i/i_0$ first becomes infinite (focus).

After these preparatory remarks the ratio of the current i at x to the unmodulated current i_0 entering the modulator can easily be derived. Let us consider a group of electrons which pass x between t and $t + dt$ and there give rise to a current i, *i.e.* have a charge $i\,|dt|$. If $x \leqq v_0 T \xi_f$ the group in question is made up exclusively of electrons which have left the modulator in the same time interval $(t_1, t_1 + dt_1)$ and which there had the charge $i_0|dt_1|$. Since no electrons have been lost en route

$$i\,|dt| = i_0\,|dt_1| \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

If on the other hand $x > v_0 T \xi_f$ the group of electrons in question is actually the sum of different groups, in the case of fig. 5 three groups which left the modulator in three different time intervals $(t_1^{(1)}, t_1^{(1)} + dt_1^{(1)}, t_1^{(2)}, t_1^{(2)} + dt_1^{(2)})$ and $(t_1^{(3)}, t_1^{(3)} + dt_1^{(3)})$.

Equation (7) must accordingly be replaced by

$$i\,|dt| = i_0 \sum_{k=1}^{3} |dt_1^{(k)}| \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

We now introduce the reduced variables (4) again. Moreover, in order not to have to distinguish each time between the cases (7) and (8) we shall collect the two cases into one equation:

$$i\,|d\tau| = i_0 \sum |d\tau_1| \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

on the understanding that here and in the equations following from it the sign of summation can be omitted when $\xi < \xi_f$.

$$i = i_0 \sum \frac{1}{|d\tau/d\tau_1|} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

If one substitutes here the value of $d\tau/d\tau_1$, which can easily be calculated from (6), the result is:

$$i/i_0 = \sum \frac{1}{|1 - \pi a \xi\,(1 + a \sin 2\pi\,\tau_1)^{-3/2} \cos 2\pi\,\tau_1|} \cdot \quad (11)$$

Equation (11) gives i/i_0 as a function of τ_1, but we wish to know i/i_0 as a function of τ and must therefore eliminate τ_1 from (11) and (6), which, however, is not possible explicitly. Figures 3 and 4 are calculated by determining corresponding values of i/i_0 and τ for a number of values of τ_1 or ξ from (6) and (11).

Figs. 2 to 5 can also be derived directly from the surface shown in *fig. 6*, which represents τ as a function of τ_1 and ξ according to equation (6). For the sake of clearness a is there assumed to be equal to 0.4. The cross-sections perpendicular to the τ_1 axis are straight lines which satisfy (6). If these straight lines are projected on the τ-ξ surface fig. 2 is obtained, while fig. 3 appears when the derivate $d\tau_1/d\tau = i/i_0$ from fig. 6 is determined and plotted as a function of τ and ξ. Fig. 4 is a cross-section of fig. 3 for τ constant, while the curves in fig. 5 are cross-sections of fig. 6 for ξ constant.

We now return to the significance of the above-mentioned critical value ξ_f of $\xi = x/v_0T$. According to equation (10) the current i at x becomes infinite [5]) when $d\tau/d\tau_1 = 0$, *i.e.* when the curve corresponding to $\xi = x/v_0T$ in fig. 5 is parallel to the τ_1 axis. It is clear from the figure that for $\xi < \xi_f$ this is never the case, for $\xi = \xi_f$ once per period and for $\xi > \xi_f$ twice per period. The critical value ξ is thus the smallest value of ξ for which the current becomes infinite. For larger values of ξ it even becomes infinite twice per period, corresponding to the double peaks in figs. 3 and 4. For small values of ξ on the other hand it always remains finite and from fig. 3 it may be seen that for small values of ξ, *i.e.*

[5]) The current maxima in question remain finite in a theory which takes the space charge into account.

close to the modulator, there are in fact no infinite peaks.

The spot in the valve for which $\xi = \xi_f$ is called the "focus", because at that point (in connection with the fact that the point for which $d\tau/d\tau_1 = 0$ is in this case at the same time a point of inflexion) the highest concentration of electrons occurs.

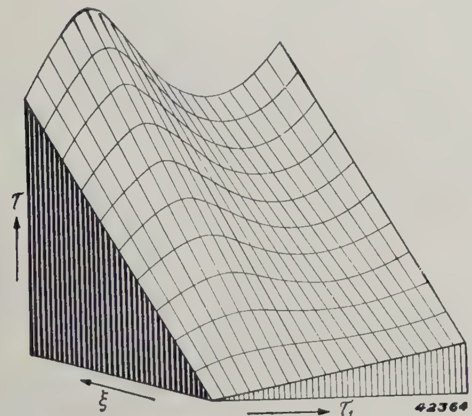

Fig. 6. The τ_1 τ_1, ξ plane (*cf.* fig. 5). The scale for τ_1 is five times as large as that for τ; $a = 0.4$.

This concentration of the electrons shows a strong analogy to the focussing of light rays by a lens and has therefore been called "phase focussing" by Brüche and Recknagel [6]). In the optical case rays of originally the same direction are focussed by a lens at a point in space; in the case of the velocity modulator electrons of originally equal velocities are focussed by an alternating field at a point in time (phase point). In the case of the lens the changes in direction of the rays depend upon the place, where they strike the lens, while in the case of the alternating field the changes in velocity of the electrons are determined by the time, or in other words by the phase at which they enter the field. The analogy is also clear from fig. 2, where the originally parallel distance-time lines (at the bottom of the figure) are broken by the A.C. field and brought to a "focus". If in fig. 2 the time axis is considered as a distance axis the case is the same as that of optical focussing.

In order to calculate ξ_f explicity the smallest value of ξ must be sought for which $d\tau/d\tau_1 = 0$, *i.e.* for which one of the denominators in (11) becomes zero. The values of ξ for which this is the case are given by

$$\frac{(1 + a \sin 2\pi\tau_1)^{3/2}}{\cos 2\pi\tau_1} \quad . \; . \; . \; . \quad (12)$$

The minimum of this is found by differentiation with respect to τ_1. For $\tau \ll 1$ the following can be derived from (12):

$$\xi_f = 1/\pi a \quad (\text{at } a \ll 1); \; . \; . \; . \; . \; . \quad (13)$$

the point of inflection then lies at $\tau_1 = 0$. For larger values of a, ξ_f becomes smaller than $1/\pi a$ In fig. 5, where $a = 0.2$, equation (13) is fairly well satisfied by $\xi_f = 1.52$ instead of $1/0.2\,\pi = 1.59$.

Calculation of the fundamental component of the current variations

Although, as stated above, it is impossible to find an explicit expression for the current variations as a function of time and place, it is possible to calculate the current value for the fundamental frequency of the current, thus the Fourier component with the frequency ω, at least if we confine ourselves to the case just considered where $a \ll 1$.

This component is usually the most important in practice, because the circuit in which it is desired to cause oscillations at the output end is tuned to the fundamental frequency and possesses a large impedance only for that frequency.

For $a \ll 1$ (6) and (11) can be developed in powers of a and the linear terms in a are sufficient. One then obtains [7])

$$2\pi\,(\tau - \xi) = 2\pi\tau_1 - \pi a\,\xi \sin 2\pi\,\tau_1 \quad . \; . \quad (14)$$

and

$$i/i_0 = \Sigma\, 1/(1 - \pi a \xi \cos 2\pi\tau_1) \; . \; . \; . \; . \; . \quad (15)$$

In connection with equation (13) the summation symbol in (15) may be omitted for $\pi a \xi \leq 1$. Calculation shows that the fundamental component of the A.C. part of i can be represented, for $\pi a \xi \leq 1$ as well as for $\pi a \xi > 1$, by [8])

$$i_2\,(\xi, \tau) = 2 i_0\, J_1\,(\pi a \xi) \cos 2\pi\,(\tau - \xi). \; (a \ll 1) \quad (16)$$

J_1 is here the Bessel function of the first order. The amplitude I_2 of the fundamental component at the position ξ thus has the value

$$I_2\,(\xi) = 2\, i_0\, J_1\,(\pi a \xi) \; . \; . \; . \; . \; . \quad (17)$$

From (17) it follows that for $a < 1$ the amplitude I_2 of the A.C. depends only on $a\xi$ and no longer on a alone, *i.e.* with varying a the same value of I_2 is retained when the distance covered by the electrons varies inversely proportional to a. For the focus ($\pi a \xi = 1$) I_2 is equal to $2\, i_0\, J_1\,(1) = 0.88\, i_0$; the maximum amplitude, however, is reached for

[6]) E. Brüche and H. Recknagel Z. Phys. **108**, 459, 1938.

[7]) When the summation sign is omitted in (15), for $\pi a \xi < 1$, these are the equations of an extended cycloid with $2\pi\,(\tau - \xi)$ and the reciprocal of i/i_0, *i.e.* i_0/i, as coordinates and τ_1 as variable parameter. For $\pi a \xi > 1$ the equation is obtained of a shortened cycloid; for certain values of τ the quantity $1/(1 - \pi a \xi \cos 2\pi\tau_1)$ then has more than one value and in order to calculate the current equation (15) must be used with the sign of summation.

[8]) Equation (14) is identical with Keppler's equation for the movement of the planets, from which Bessel solved τ_1 as a function of τ with the aid of the functions which were later named after him. The solution of $d\tau_1/d\tau = i/i_0$ as a function of τ leads to (16).

$\pi a\xi = 1.84$ and amounts to $2\, i_0\, J_1\,(1.84) = 1.16\, i_0$. The plane in which this happens we call the "maximum plane", to distinguish it from the focus; the following holds for that plane:

$$\pi(\xi a)_m = 1.84\,, \quad . \; . \; . \; . \; . \quad (18a)$$

Hence if a is taken as constant:

$$\xi_m = 1.84/\pi a. \quad . \; . \; . \; . \; . \; . \quad (18b)$$

The velocity modulator as amplifier

Before considering the velocity modulator as an exciter of oscillations we shall first study its action as an amplifier. For that purpose an A.C. voltage $aV_0 \sin \omega t$ is applied between the grids B_1 and A_1 (modulator). The energy of the beam of electrons modulated by this A.C. voltage is given off to an oscillation circuit with impedance Z tuned to ω and connected between A_2 and B_2 (inductor). We shall now assume that the distance d_2 between A_2 and B_2 is very small as compared with the concentration wave length λ_i. Then the A.C. which flows from B_2 through the impedance Z to A_2 is exactly equal to the value i_2 in (16) of the fundamental component of the current through the valve with the amplitude I_2 [9]). If V_2 is the amplitude of the A.C. voltage between B_2 and A_2 and the impedance for the higher harmonics may be ignored, compared with Z, the following holds: $V_2 = I_2 Z$, and the total oscillation energy which flows through the output circuit per second is $^1\!/_2\, I_2 V_2$. If, further, we understand by the efficiency η the ratio of this total oscillation energy to the D. C. energy $i_0 V_0$ given off to the electrons, then according to (17)

$$\eta = \frac{I_2\, V_2}{2 i_0 V_0} = \frac{V_2}{V_0}\, J_1\,(\pi a \xi) \quad . \; . \; . \quad (19)$$

In order to give this efficiency its maximum value two conditions must be fulfilled, namely $I_2/2\, i_0$ or $J_1\,(\pi a\xi)$ must be a maximum and V_2/V_0 must be a maximum. The first condition implies that the maximum plane must be situated at the position of the inductor; the corresponding value ξ_m of ξ follows from (18b), while

$$I_2/2 i_0 = J_1\,(\pi a \xi_m) = J_1\,(1.84) = 0.58\,. \quad (20)$$

As to the second condition it may be noted that for V_2/V_0 electrons will be thrown back before they reach the inductor. We thus assume that V_2 can at the most be equal to V_0. It then follows from (19) that

$$\eta_{\max} = 0.58 \quad . \; . \; . \; . \; . \; . \quad (21)$$

The amplification factor, which is obviously given by V_2/aV_0, is therefore in the case of maximum efficiency equal to the reciprocal $1/a$ of the depth of modulation.

The conditions which must be satisfied by i_0, V_0 and Z at given values of the depth of modulation a and length of the overtake space in order to attain this maximum efficiency follow from (5) and the relation $V_2 = V_0$.

According to (5) the following holds:

$$\sqrt{V_0} = 500\; x/\lambda\xi \quad (V_0 \text{ in volt})\,, \quad . \quad (22)$$

thus, on the basis of (18)

$$\sqrt{V_0} = 855\; ad/\lambda\; (V_0 \text{ in volt}) \quad . \; . \; . \quad (23)$$

Further, for the realization of $\eta_{\max}$, Z must have a value Z_m which satisfies $V_2 = 1.16\; i_0 Z_m = V_0$, thus $Z_m = V_0/1.16\; i_0$.

If we denote V_0/i_0, the "valve resistance", by Z_b and the ratio of external resistance Z to valve resistance Z_b by ϱ, then

$$\varrho = Z/Z_b = Z\, i_0/V_0 \quad . \; . \; . \; . \; . \quad (24)$$

$$\varrho_m = Z_m/Z_b = 1/1.16 = 0.86\,. \; . \quad (25)$$

At a depth of modulation a, therefore, theoretically the maximum efficiency of 58% is obtained at $\sqrt{V_0} = 855\; ad/\lambda$ and $\varphi = \varrho_m = 0.86$.

The velocity modulator as oscillator

In order to use the velocity modulator as oscillator there must be a back-coupling of the inductor to the modulator. Just as in the case of triode oscillators, this back-coupling must have a certain minimum value to make the occurrence of oscillations possible. The "back-coupling condition" can easily be indicated, after the foregoing.

When an A.C. voltage $V_1 \sin 2\pi\tau$ is applied between B_1 and A_1 an A.C. occurs in the impedance of the inductor, whose fundamental frequency i_2 is determined by (16), so that for the voltage over the impedanze Z between B_2 and A_2 the following is valid:

$$-i_2\, Z = -2\, i_0 Z\, J_1\,(\pi a\xi) \cos 2\pi\,(\tau - \xi) = \\ = -V_2 \cos 2\pi\,(\tau - \xi)\,. \quad (26)$$

The amplitude of this A.C. voltage thus amounts to

$$V_2 = 2 i_0\; Z\; J_1\,(\pi a \xi) \quad . \; . \; . \; . \; . \quad (27)$$

If it is desired to maintain oscillations a part K (back-coupling factor) of V_2 must be sent back to

[9]) See for example C. J. Bakker and G. de Vries, Physica, 2, 683, 1935, where the case for finite $d = d_2/\lambda_1$ is discussed. For this the amplitude becomes

$$2\, i_0\, \frac{\sin \pi\tau}{\pi\tau}\, J_1\,(\pi a \xi).$$

A_1B_1 and must there be equal to V_1 in amplitude and phase. We shall return to the phase condition later. The amplitude condition obviously becomes

$$KV_2 = V_1 = aV_0, \quad . \quad . \quad . \quad . \quad (28)$$

or in connection with (27) and (24)

$$K \cdot 2i_0 \, Z \, J_1 \, (\pi a \xi) = aV_0,$$

so that

$$K = a/2\varrho J_1(\pi a \xi) \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

At the beginning of the oscillations when a is practically equal to zero, K must have a value $K_{\min}$ which according to (29) satisfies

$$K_{\min} = 1/\pi \varrho \xi, \quad . \quad . \quad . \quad . \quad . \quad (30)$$

because for $a \ll 1$, $2 \, J_1 \, (\pi a \xi) = \pi a \xi$.

When one takes $K > K_{\min}$ the amplitude of the oscillations increases; the value of the depth of modulation a in the modulator then follows from (29).

The efficiency is again determined by (19), for which, with the relation $V_2/V_0 = a/K$ following from (28), one may also write

$$\eta = \frac{I_2}{2i_0} \frac{V_2}{V_0} = \frac{a}{K} J_1 \, (\pi a \xi) \quad . \quad . \quad . \quad (31)$$

Just as in the case of the amplifier, η is a maximum when $I_2/2i_0$ and V_2/V_0 or a/K are maxima. The latter is again true for $V_2 = V_0$ or $a = K$. The conditions for maximum efficiency (18*a*) and (25) hereby pass over into:

$$\left.\begin{array}{l} \pi(\xi K)_m = 1.84 \quad (32) \\ \varrho_m = 0.86 \quad (33) \end{array}\right\} \eta_{\max} = 0.58.$$

Equations (29) and (31), however, lead not only to the conditions for $\eta_{\max}$ but also give the efficiency for any other case, since it depends upon the two characteristic quantities ϱ and ξK. This may be seen more clearly when (29) and (31) are written in the following form:

$$\varrho \xi K = a\xi/2J_1(\pi a \xi) \quad . \quad . \quad . \quad . \quad . \quad (34)$$

$$\eta \xi K = a\xi J_1(\pi a \xi) \quad . \quad . \quad . \quad . \quad . \quad . \quad (35)$$

From equation (34) the value of $a\xi$ at given values of ϱ and ξK follows, while substitution of this in (35) gives the efficiency. It is also possible to calculate both $\varrho \xi K$ and $\eta \xi K$ as functions of $a\xi$ according to (34) and (35) and then represent $\eta \xi K$ as a function of $\varrho \xi K$ in a single curve.

In *fig. 7* η is plotted as a function of ξK with ϱ as a parameter. At a constant value of ϱ the maximum of η is reached when the inductor lies in the maximum plane. For $\varrho = 0.86$ this maximum reaches its optimum value of 58%. For $\varrho > 0.86$ only those parts of the curve are drawn for which $V_2 < V_0$, *i.e.* $a < K$, since with $V_2 > V_0$ electrons would be reflected. It may be seen from the figure that for these values of ϱ the efficiency always remains 58%.

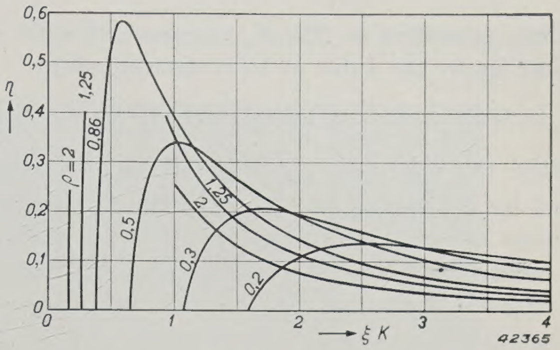

Fig. 7. Theoretical value of the efficiency η of a velocity-modulator valve as a function of ξK at constant ϱ. K = back-coupling factor, ξ = number of parcels between the modulator and the inductor, ϱ = ratio of external impedance Z to valve resistance V_0/i_0. The theoretical optimum efficiency of 58% is attained at the maximum of the curve for $\varrho = 0.86$.

Fig. 7 may also be interpreted as follows: for $\varrho < 0.86$ the current through the valve can be entirely modulated by taking $\pi a \xi = 1.84$; the impedance, however, is then still too small to modulate also the voltage to its maximum value ($V_2 = V_0$). This latter takes place only at $\varrho = 0.86$. For $\varrho > 0.86$ on the other hand the impedance is so large that the voltage can be entirely modulated; this takes place at constant ϱ for two different values of ξK and $a\xi$ respectively (in the figure the points where the curves end). For these values of ϱ, however, the current at the inductor is again not entirely modulated, so that η remains smaller than 58%.

According to (35) the maxima of η for $\varrho < 0.86$ (current maximum) in fig. 7 lie on the equilateral hyperbola $\pi \eta \xi K = 1.84 \, J_1 \, (1.84)$. For $\varrho > 0.86$ the curves end at the points for which $V_2 = V_0$ (voltage maximum) or $a = K$; these points, according to (35), lie on the curve $\xi = J_1 \, (\pi \xi K)$ and according to (34) satisfy the relation $2\varrho \, J_1 \, (\pi \xi K) = 2\varrho \eta = 1$.

In the foregoing no use has been made of the phase condition of the back-coupling. According to (26) the voltage between B_2 and A_2 is

$$-V_2 \cos 2\pi(\tau - \xi) = V_2 \sin 2\pi(\tau - \xi - {}^1/_4), \quad (36)$$

while that between B_2 and A_2 was taken equal to

$$V_1 \sin 2\pi\tau.$$

For the phase difference Φ between the two voltages, therefore, the following holds:

$$\Phi = (\xi + {}^1/_4) \, 2\pi.$$

Due to the back-coupling the phase is rotated through an angle φ. The phase condition of the back-coupling is now apparently:

$$\Phi + \varphi = (\xi + 1/4)\, 2\pi + \varphi = n \cdot 2\pi \text{ (} n \text{ a whole number)},$$

$$\xi = n - 1/4 - \varphi/2\pi \quad . \quad . \quad . \quad . \quad (37)$$

Further, according to (22) V_0 is connected with ξ, so that finally the value of V_0 is determined by

$$\sqrt{V_0} = 500\, d/\lambda\, (n - 1/4 - \varphi/2\pi)\,. \quad (V_0 \text{ in volts}) \quad (38)$$

When the values of φ, d, λ, K and Z are determined by the oscillator system (modulator circuit, inductor circuit and back-coupling connections), the values of V_0 and i_0 may still be chosen. For the excitation of oscillations ξ must satisfy (37). Hereby several discrete values of ξ are fixed, and thus, in connection with (38), also of V_0. The value of i_0 can then still be chosen arbitrarily. This determines $\varrho = i_0\, Z/V_0$. The efficiency, which is a function of ϱ and ξK, then follows from fig. 7.

We have defined the efficiency η as the ratio of the total oscillation energy to the energy $i_0 V_0$ given off to the electrons. In practice the "aerial efficiency" η_a is of more importance. By this we understand the quotient of the aerial power and $i_0 V_0$. Obviously $\eta = \eta_a + \eta_s$ where η_s refers to the oscillator system.

If now the impedances Z_a and Z_s observed from A_2B_2 (fig. 1) of the aerial and the oscillator system respectively are known, the following relation holds for the total impedance Z: $1/Z = 1/Z_a + 1/Z_s$; ϱ is then known, and from fig. 7 η and, after some calculation, η_a also can be derived.

In velocity modulators η_a is often appreciably smaller than η, since Z_s cannot be taken large enough. Usually one works here with high voltages and currents whereby V_0/i_0 becomes high and ϱ low. As an example we choose $V_0 = 5$ kV, $i_0 = 50$ mA and $\xi K = 1$. Then $Z_b = 100\,000\ \Omega$. For the occurrence of oscillations according to (30) $\varrho\,\xi K$ must at least be equal to $1/\pi$, thus ϱ at least $1/\pi$ and Z at least $32\,000\ \Omega$. If Z_s is little larger than this amount, Z_a must be taken very large, since otherwise the total Z becomes too small and the oscillations stop. The aerial can then take up only little energy and η_a will be much smaller than the value of η without aerial (η_0) following from fig. 7. Thus for example for $Z_s = 40\,000\ \Omega$, for the case with no aerial $\eta_0 = 0.22$, while with an aerial the highest value attainable for η_a is only 0.01. With increasing Z_s the ratio η_a/η_0 becomes more favourable; for $Z_s = 100\,000\ \Omega$ thus $\varrho = 1$, we arrive at $\eta_0 = 0.38$ (without aerial) and $\eta_a = 0.17$ maximum. For $Z_s = \infty$ all the energy can be concentrated in the aerial.

Finally it must be remarked that in comparing the experimental results with the results of the above theory no precise agreement can be expected, since the theory is derived on the following simplifying assumptions:

1) the space charge may be disregarded;
2) the depth of modulation is small ($\alpha \ll 1$);
3) the grids are completely permeable for electrons;
4) the electrons always move parallel to the axis;
5) the grid distances are infinitesimally small.

As far as point 4) is concerned it may be noted that when slits with no grids are used (see below) the field is very unhomogeneous and, moreover, the magnetic field holding the electron beam together causes the electrons to describe spiral paths. This may cause a flattening of the maxima along the electron beam.

Empirical example

In order to elucidate the above theoretical considerations we shall discuss an experimental case which has indeed only a small efficiency but in which the various characteristic quantities are easy to determine. The arrangement is shown in *fig. 8*. The beam of electrons here passes through a tube made of silica to reduce the dielectric losses. With the aid of a coil, not shown in the figure, the beam of electrons is kept concentrated along the axis of the tube. The oscillation system, shown in cross-section for the sake of clarity, consists of a co-axial Lecher system of copper outside the silica tube; the inner and outer cylinders are kept at the proper distance from each other by two insulating discs.

In the silica tube itself there are no grids of any kind for varying the velocities of the electrons, the beam being affected only by external electrodes. In the left-hand slit the electron velocity is modulated, in the inner cylinder the velocity modulation is converted into a density modulation and this density modulation induces again an A.C. field in the right-hand slit. As a result the whole system oscillates in such a way that voltage maxima occur at the slits, while the length of the inner cylinder is for example about $^1/_2\ \lambda$. The energy can be taken off by a loop, which is brought into the interior through a slit, which may be seen at the middle of the upper side of the outer cylinder.

Instead of the grids A_1B_1 and A_2B_2 of fig. 1 we here have the slits S_1T_1 and S_2T_2 (*fig. 9*). The impedance Z between S_2 and T_2 is approximately equal to the impedance between S_2 and U_2 of the open $^1/_2\ \lambda$ system $T_1\ U_1\ U_2\ S_2$ and can be calculated by means of the formula [10])

$$Z = \frac{240}{\pi\delta}\,\frac{R_1R_2}{R_1 + R_2}\left(\ln\frac{R_2}{R_1}\right)^2,$$

where R_2, the radius of the outer cylinder, equals 3 cm, R_1, that of the inner cylinder, equals 0.95 cm and δ, the depth of penetration of the skin effect,

[10]) See C. G. A. von Lindern and G. de Vries Lecher systems, Philips Techn. Rev. 6, 240, 1941.

is $4\times10^{-5}\sqrt{\lambda}$ (copper). With $\lambda = 35$ cm in the case in question this gives approximately

$$Z = 3 \times 10^6\ \Omega.$$

The inside of the cylinder $T_1\ S_2$ (length $d = 15$ cm) here serves as overtake space, while the outside functions as back-coupling connection of about $^1/_2\lambda$ length. Since T_1 and S_2 are opposite in phase, According to (38) the tube voltage is thereby fixed:

$$\sqrt{V_0} \approx 250/(n-{}^1/_4),\ (V_0 \text{ in volts}) \qquad (39)$$

since here $d \sim {}^1/_2\ \lambda$. For the back-coupling factor K the following holds:

$$K = 1.$$

The valve oscillates at $V_0 = 5000$ volts, for which

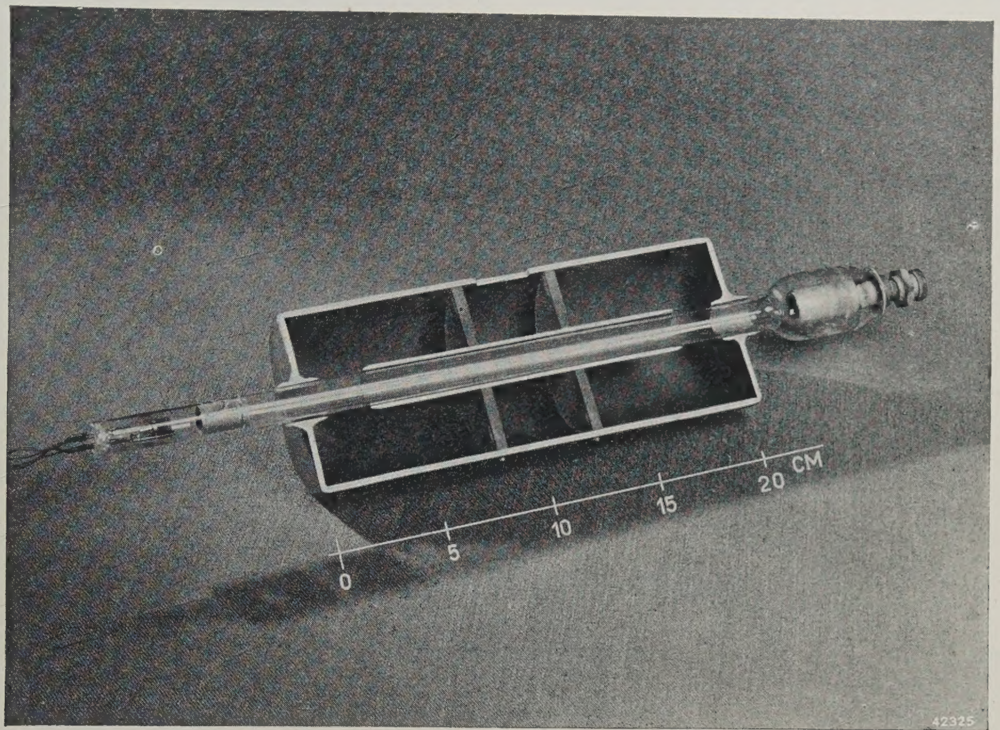

Fig. 8. External oscillator system consisting of two co-axial cylinders (shown in cross-section) with two slits. The oscillations are excited by the electron beam in a silica tube inserted along the axis of the oscillator system. The whole is placed in a co-axial magnetic field.

the angle φ, through which the phase is rotated upon back-coupling, is zero, so that according to (37) the number of parcels in the length d becomes

$$\xi = n - {}^1/_4.$$

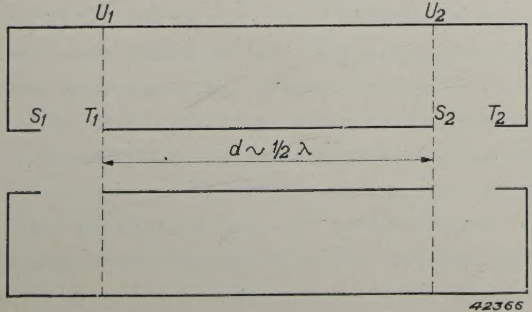

Fig. 9. Diagram of the oscillator system of fig. 8. The system oscillates in about the same way as the open Lecher system $T_1\ U_1\ U_2\ S_2$ with a length of $^1/_2\ \lambda$. The outside of the cylinder $T_1\ S_2$, inside of which is the overtake space, serves as back-coupling connection with $K = 1$.

according to (39) n is approximately equal to 4, so that $3^3/_4$ parcels are situated within the overtake space [11]).

According to (30) the oscillations begin at $\varrho\xi K = 1/\pi$. The width of the slit here, however, is not infinitesimal, which makes the minimum value of $\varrho\xi K$ larger (*cf.* footnote 9)). When this is taken into account, with $\xi = 3^3/_4$, $V_0 = 5000$ volts and $Z = 3 \times 10^5\ \Omega$, one finds for the current at which the oscillations begin a value of 2 mA. The experimental value was $3^1/_2$ mA.

Instead of having the oscillator system external it may also form part of the valve itself. In that way oscillations of considerably higher frequency can be excited.

11) With an external oscillator system deviations from equation (39) may occur, since due to wall charges the energy of the electrons need not correspond to the anode voltage V_0.

It is clear that with the system of fig. 8 only a low efficiency can be expected, since according to fig. 7 when $\xi \approx 4$ and $K = 1$, thus $\xi K = 4$, the total oscillation efficiency η is not more than 10%, only part of which reaches the aerial. Experimentally in the case of fig. 8 a power of 8 watts could be realized in an incandescent lamp, which corresponds to an efficiency of $\eta_a = 3\%$. In judging this result account must be taken of the simplifications introduced into the theory (see above).

In order to obtain the theoretical maximum efficiency of 58%, according to (32) $\pi\xi K$ should be taken equal to 1.84. On the other hand for $K = 1$ ξ must equal $n - {}^1/_4$ (n a whole number); the optimum value of ξ is thus ${}^3/_4$. This leads, however, in connection with (39), to an unusable value for V_0 ($>$ 100 000 volts). Therefore in order to attain high efficiencies the oscillator systems must be designed otherwise, but we shall not go more deeply into that here.